ACTIVE R BANDPASS FILTER

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for the Degree of
MASTER OF TECHNOLOGY

By
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to the
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CERTIFICATE

This is to certify that the thesis entitled "Active R Bandpass Filter" is a record of work carried out under my supervision and that it has not been submitted elsewhere for a degree.

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October 1978. S.S.Pujari

ABSTRACT

The historical developments of Active R filter is given and some early works are discussed. The proposed bandpass circuits, using Op-amp pole, are developed. Op-amp μA741C has been used which was modelled to have a single pole upto 10 times the operating frequency. The experimental verification of the bandpass filter is carried out and compared with the theoretical results. Conclusion is drawn. The future applications and problems of Active R filters are discussed.

CONTENTS

LIST OF SYMBOLS

A_o	=	Open loop d.c. gain of Op-amp
$1/\tau$	=	Open loop first pole in radians/sec.
B	=	A_o/τ = Gain bandwidth product
s	=	Laplace transform
ω_o	=	Centre frequency in radians/sec
Q	=	Quality factor
G_o	=	Midband gain
V	=	Power supply in volts
T	=	Temperature in °C.

LIST OF FIGURES

CHAPTER 1

INTRODUCTION

1.1 Early Works

Historically, the parasitic capacitances associated with active devices have been considered undesirable and much attention has been given to compensating for their effects. Recently, Allen and Means [1] have suggested that these parasitic capacitance might be made full use of in order to design both grounded and semi-floating inductors without the need of capacitors even.

Even before this, Caparelli [2] suggested the use of the lowpass properties of the transistors for the design of active filters without capacitors. This work was extended in a series of papers by Caparelli and Libertore [3] to apply to a number of specific filter characteristics and was finally generalised by Berman and Newcomb [4].

In an analogous manner, Rao and Srinivasan [5] have suggested that the pole of an operational amplifier could be used to design active filters, with one Op-amp and grounded capacitor. This is the first pioneering work done in this field. This was followed by a series of works by different authors. A.K. Mitra and V.K.Atre [6], Soliman and Fawzy [7], R. Nandi [8], have used one Op-amp and one capacitor to realise bandpass filter. Recently R. Nandi [9] has realised bandpass or lowpass filter, using one Op-amp and one capacitor.

Rao and Srinivasan [10] realised bandpass filter with only two Op-amps and completely eliminated capacitor. Srinivasan [11], Schawman [12], A.K. Mitra and V.K. Atre [13], and Li and Li [14] have realised bandpass and lowpass filter using only two Op-amps. Soderstand [15] improved Schauman's circuit [12] and realised a two Op-amp circuit which has LP, BP and HP transfer characteristics. He adds another Op-amp, considered ideal, for a BE filter. Soliman and Fawzy [16] proposed a three Op-amps universal filter which has LP, BP and biquadratic, non-minimum phase transfer characteristics. Soderstand [17] reduced loading by going over to a CMOS version of his earlier circuit.

Venkateswaran and Sowrirajan (1978) has applied the concept of driving point function in obtaining the transfer function of a BP filter using OP-amp pole. Venkateswaran (1978) proposed a multifunction active R filter with BP, LP and BE transfer function.

In the present paper second order bandpass filter has been realised using two and three Op-amps separately. This paper is based on the analogous Op-amp version of the transistorised bandpass filter realised by Caparelli and Libertore [3].

1.2 Active R Filter Realised by One OP-amp and One Capacitor

1.2.1 Rao, K.R. and Srinivasan, S. [5] :

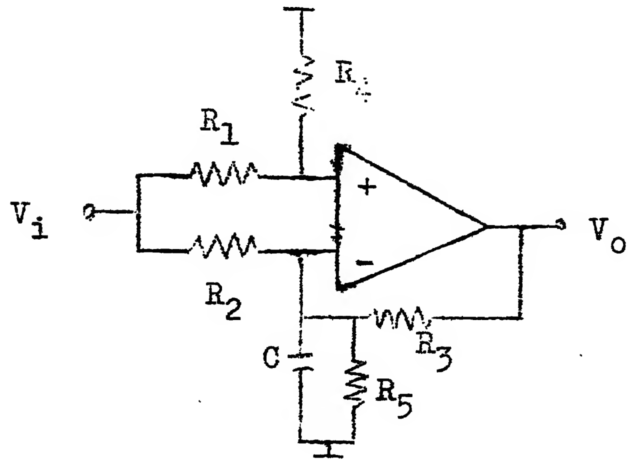


Fig.2.1

$$T_{BP} = \frac{V_o}{V_i} = \frac{(SCR + \frac{R}{R_5})/2}{S^2(\frac{RC}{B}) + \frac{S}{B}(2 + \frac{R}{R_5})+1} \tag{1.1}$$

$$\omega_o = \left(\frac{B}{CR_3}\right)^{\frac{1}{2}} \tag{1.2}$$

$$Q = \frac{1}{K}(BCR_3)^{\frac{1}{2}} \tag{1.3}$$

$$G_o = \frac{B}{2K}(CR_3 + \frac{1}{SR_5}) \tag{1.4}$$

where $R_1 = R_2 = R_3 = R_4 = R$, $K = 2 + \frac{R}{R_5}$.

Comments: Equation (1.1), shows that the numerator contains a constant term, which is not an exact bandpass function unless $R_5 = \infty$. ω_o and Q are independently tunable but not G_o. Q and G_o variations with the supply voltage V, can be minimised by proper choice of

CR product for a particular frequency. ω_o variation with respect to V can be compensated around any ω_o by introducing a voltage dependent variation in either C or R in exactly the same manner as $B^{\frac{1}{2}}$ varies with V. The voltage dependent output resistance r_d of a JFET has been introduced as a part of R_3, and sensitivity figure was found to be reduced. A similar compensation scheme for temperature variations of ω_o can be incorporated knowing the temperature variations of A_o, τ and B. This scheme will use a thermistor in place of the JFET.

1.2.2 Mitra, A.K. and Atre, V.K.[6]:

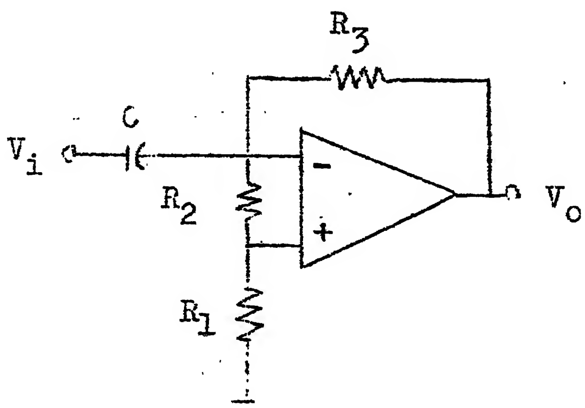


Fig. 1.2

$$T_{BP} = \frac{V_o}{V_i} = \frac{SCR_3\left(\frac{R_2}{R_1+R_2}\right)B}{S^2 + \frac{S}{CR_3} + \frac{1}{CR_3}\left(\frac{R_2}{R_1+R_2}\right)B} \tag{1.5}$$

$$\omega_o = \frac{1}{CR_3}\left(\frac{R_2}{R_1+R_2}\right)B \qquad (1.6)$$

$$Q = CR_3\left(\frac{R_2}{R_1+R_2}\right)B \qquad (1.7)$$

$$G_o = C^2R_3^2\left(\frac{R_2}{R_1+R_2}\right)B \qquad (1.8)$$

<u>Comments</u>: Independent control of ω_o and Q is possible. The sensitivities are less than 1 for ~~medium~~ Q factors and high frequencies.

1.2.3 Mitra, A.K. and Atre, V.K. [6]:

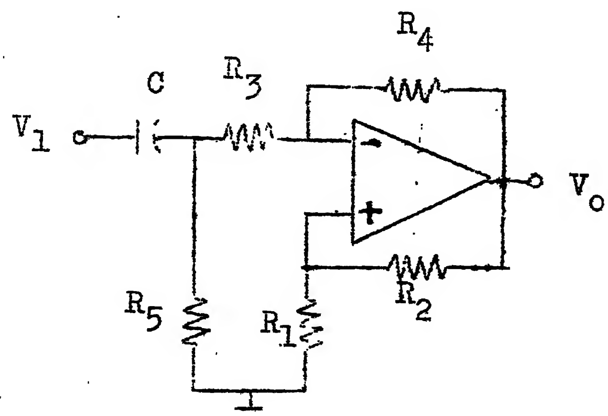


Fig. 1.3

$T_{BP} = V_o/V_1$

$$= \frac{-SB}{S^2 + S\,\frac{1}{C}\left(\frac{R_4+R_5}{R_4R_5}\right) + \frac{B}{CR_4}} \qquad (1.9)$$

Assuming $R_3/R_4 = R_1/R_2$ and $R_2 >> R_1$

$$\omega_o = \left(\frac{B}{CR_4}\right)^{\frac{1}{2}} \qquad (1.10)$$

$$Q = \frac{B}{\omega_o}\left(\frac{R_5}{R_4+R_5}\right) \tag{1.11}$$

$$G_o = \left(\frac{R_4+R_5}{R_5}\right) Q^2 \tag{1.12}$$

<u>Comments</u>: Independent control of ω_o and Q is possible. The circuit can be tuned for a higher Q by making $R_1/R_2 >> R_3/R_4$. The Q factor varied from 13.12 to 46.0 (by varying R_1) and f_o changed by 3.24% and G_o varied from 44.6 to 55.6 dB.

1.2.4 Soliman and Fawzy [7] :

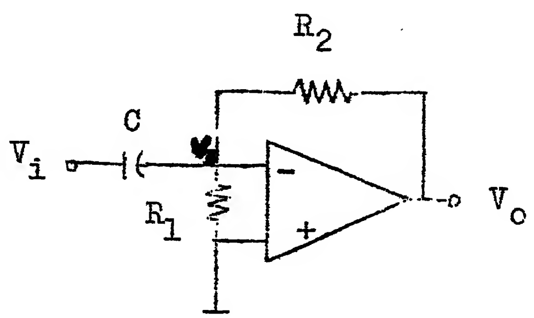


Fig.1.4

$$T_{BP} = \frac{V_o}{V_i} = -\frac{SB}{S^2+S\,\frac{1}{C}\left(\frac{R_1+R_2}{R_1R_2}\right) + \frac{B}{CR_2}} \tag{1.13}$$

$$\omega_o = \left(\frac{B}{CR_2}\right)^{\frac{1}{2}} \tag{1.14}$$

$$Q = \frac{R_1}{R_1+R_2}\,(CR_2B)^{\frac{1}{2}} \tag{1.15}$$

$$G_o = C\left(\frac{R_1R_2}{R_1+R_2}\right)B \tag{1.16}$$

$$\omega_o Q = GB\left(\frac{R_1}{R_1+R_1}\right) \tag{1.17}$$

$$T_{HP} = \frac{V_3}{V_1} = \frac{s^2}{s^2+s\,\frac{1}{C}\left(\frac{R_1+R_2}{R_1\ R_2}\right) + \frac{B}{CR_2}} \tag{1.18}$$

Comments: Expression (1.13) has no cancellation term in the numerator which is an advantage over the circuit of Rao and Srinivasan [5]. The ω_o Q product, when $R_1 = \infty$, is double that of Rao and Srinivasan circuit. Expression (1.18) is a highpass transfer function whose practical utility is limited unless the output is followed by an ideal voltage follower.

1.2.5 Nandi, R. 9 :

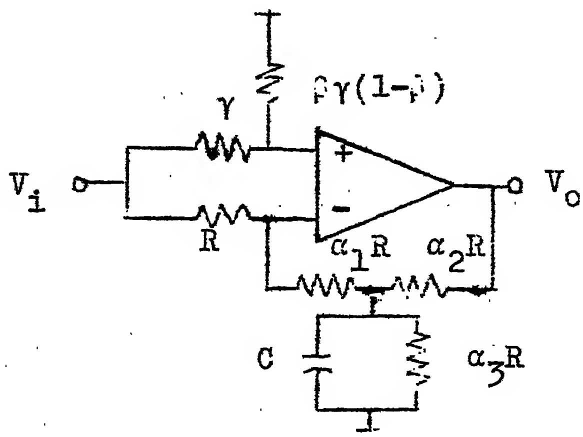


Fig. 1.5

$$\frac{V_o}{V_i} = B \frac{SRC\alpha_2\alpha_3(\beta(1+\alpha_1)-\alpha_1)+\beta(\alpha_2(1+\alpha_1+\alpha_3)+\alpha_3(1+\alpha_1)) -(\alpha_1\alpha_2+\alpha_2\alpha_3+\alpha_3\alpha_1)}{S^2RC\alpha_2\alpha_3(1+\alpha_1)+S(\alpha_2(1+\alpha_1+\alpha_3)+\alpha_3(1+\alpha_1)) + \alpha_3 B} \tag{1.19}$$

Bandpass Realisation $\beta = \dfrac{\alpha_1\alpha_2 + \alpha_2\alpha_3 + \alpha_1\alpha_3}{\alpha_2(1+\alpha_1+\alpha_3)+ \alpha_3(1+\alpha_1)}$

Lowpass Realisation $\beta = \dfrac{\alpha_1}{1 + \alpha_1}$

$$\omega_o = \left[\frac{B}{\alpha_2(1+\alpha_1)RC}\right]^{\frac{1}{2}} \tag{1.20}$$

$$Q = \frac{[\alpha_2(1+\alpha_1)RC\ B]^{\frac{1}{2}}}{1+\alpha_1+[(\alpha_2/\alpha_3)(1+\alpha_1+\alpha_3)]} \tag{1.21}$$

Bandpass midband gain $G_{om} = \dfrac{\alpha_2}{1+\alpha_1} Q^2$ (1.22)

Lowpass midband gain $G_{op} = \dfrac{\alpha_2}{1+\alpha_1}$ (1.23)

<u>Comments</u>: This is a versatile structure in that in the same circuit realizability of either bandpass or lowpass function is possible by adjustment of a single resistor.

1.3 <u>Active R Filter Realised by Two Op-amps</u>

1.3.1 Rao, Radhakrishnan [10]:

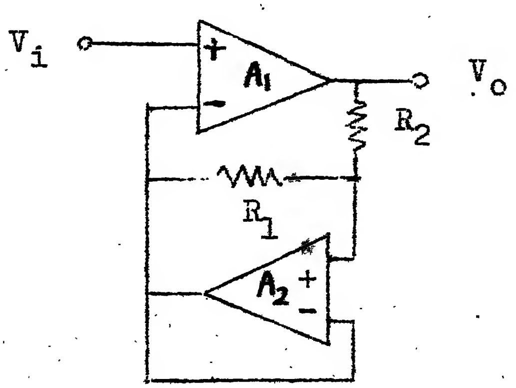


Fig.1.6

$$T_{BP} = \frac{V_o}{V_1} = \frac{SB_1 + \left(\frac{B_1B_2}{K}\right)}{S^2+S(B_2/K)+(B_1B_2/K)} \tag{1.24}$$

$$\omega_o = (B_1B_2/K)^{\frac{1}{2}} \tag{1.25}$$

$$Q = (K(B_1/B_2))^{\frac{1}{2}} \tag{1.26}$$

$$G_o = \sqrt{K^2(B_1/B_2)^2 + (B_1/\omega S)^2} \tag{1.27}$$

$$K = \frac{R_1+R_2}{R_1}$$

<u>Comments:</u> Expression (1.14) shows that the numerator has a constant term, which is not an exact bandpass function.

1.3.2 Rao Radhakrishnan 10 :

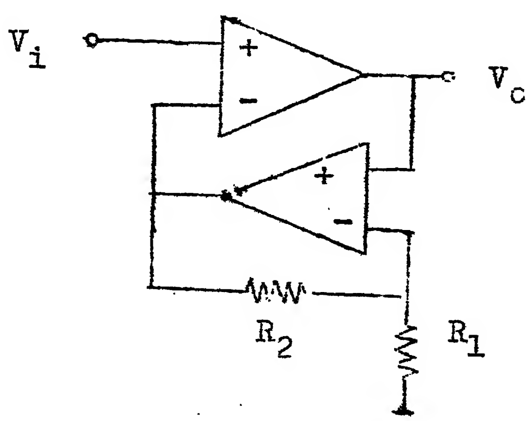


Fig.1.7

$$T_{BP} = \frac{SB_1+\left(\frac{B_1B_2}{K}\right)}{S^2+S\left(\frac{B_2}{K}\right) + (B_1B_2)} \tag{1.28}$$

$$\omega_o = (B_1B_2)^{\frac{1}{2}} \tag{1.29}$$

$$Q = K(B_1/B_2)^{\frac{1}{2}} \tag{1.30}$$

$$G_o = \sqrt{K^2\left(\frac{B_1}{B_2}\right)^2 + \left(\frac{B_1}{\omega}\right)^2} \tag{1.31}$$

Comments: The Q and G_o are insensitive to variations in the temperature and the bias voltages of the Op-amp while $_o$ varies with them. Variation in $_o$ can be accomplished by variation in B with V. An attenuator scheme is given as shown in Fig. 1.8 which changes the BW product of the amplifier.

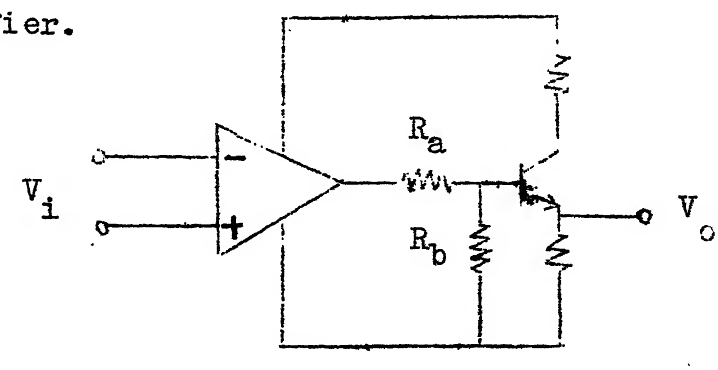


Fig.1.8

gets modified to

$B \to bB$ where $b = \frac{R_a}{R_a+R_b}$

Compensation of temperature($S_\theta^B = -S_\theta^b$) can be accomplished by introducing a thermistor in one of the resistors of the Op-amp. This has the lowest sensitivity performance if we can assume the availability of Gain Bandwidth product

stabilised Op-amps. Ghausi and Acar [19] has used this block in a FLF design to realise an active R sixth order bandpass Butterworth filter.

1.3.3 Schawman [12]:

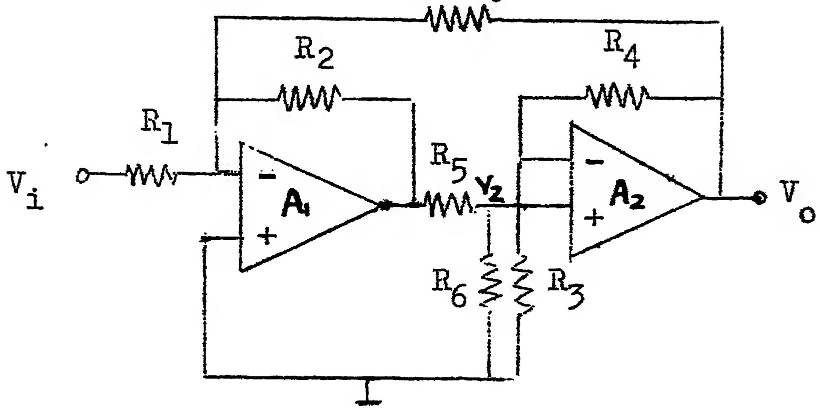


Fig 1.9

$$T_{BP} = \frac{V_2}{V_1} = \frac{-\left(S+\frac{R_3B_2}{R_3+R_4}\right)\frac{R_o}{R_1}B_1}{S^2\left(1+\frac{R_o}{R_1}+\frac{R_o}{R_2}\right)+S\left(B_1+\left(1+\frac{R_o}{R_2}+\frac{R_o}{R_1}\right)\frac{R_3}{R_3+R_4}B_2\right)+B_1B_2\left(\frac{R_3}{R_3+R_4}+\frac{R_6}{R_5+R_6}\right)} \quad (1.32)$$

$$T_{LP} = \frac{V_o}{V_1} = \frac{(-R_o/R_1)B_1B_2}{S^2\left(1+\frac{R_o}{R_1}+\frac{R_o}{R_2}\right)+S\left(B_1+\left(1+\frac{R_o}{R_2}+\frac{R_o}{R_1}\right)\frac{R_3}{R_3+R_4}B_2\right)+B_1B_2\left(\frac{R_3}{R_3+R_4}+\frac{R_6}{R_5+R_6}\right)} \quad (1.33)$$

1.3.4 Mitra, A.K. and Atre, V.K. [13]:

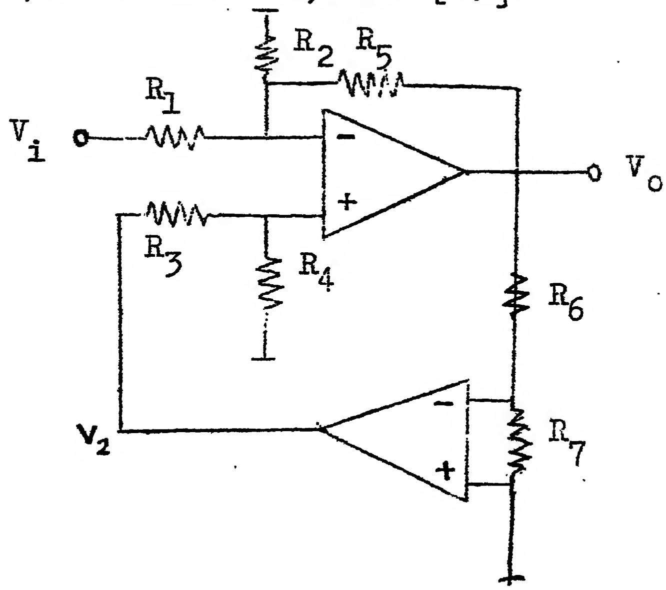


Fig.1.10

$$T_{BP} = \frac{V_o}{V_1} = \frac{-\, S\left(\dfrac{1}{\left(1+\dfrac{R_1}{R_2}+\dfrac{R_1}{R_5}\right)}\right)B_1}{D} \qquad (1.34)$$

$$T_{LP} = \frac{V_2}{V_1} = \frac{\left(\dfrac{1}{\left(1+\dfrac{R_1}{R_2}+\dfrac{R_1}{R_5}\right)}\right)B_1 B_2}{D} \qquad (1.35)$$

where

$$D = S^2 + S\left(\frac{1}{1+\dfrac{R_5}{R_1}+\dfrac{R_5}{R_2}}\right) B_1 + \left(\frac{R_3}{R_3+R_4}\right)\left(\frac{R_7}{R_6+R_7}\right)B_1 B_2$$

The scheme as shown in Fig. 1.11 has been used to get all pass and notch functions.

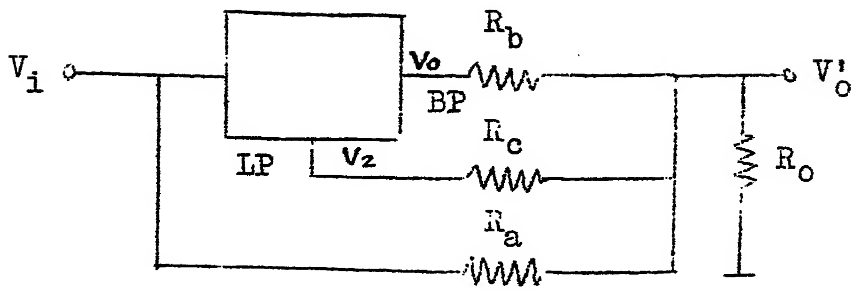


Fig.1.11

Let

$$\frac{V_2}{V_i} = \frac{k_o \omega_o^2}{S^2 + S\frac{\omega_o}{Q} + \omega_o^2}, \qquad V_o/V_i = \frac{-SH_o}{S^2 + S\frac{\omega_o}{Q} + \omega_o^2}$$

$$\frac{R_a}{R_o} = \lambda, \quad \frac{R_a}{R_b} = r, \quad \frac{R_a}{R_c} = \delta, \quad k = \frac{1}{1+r+\delta+\lambda}, \quad k' = \frac{1}{1+\lambda+r}$$

$$\frac{V_o'}{V_i} = k\,\frac{S^2 - (H_o r - \frac{\omega_o}{Q})S + \omega_o^2(1+k_o\,\delta)}{S^2 + S\frac{\omega_o}{Q} + \omega_o^2} \tag{1.36}$$

<u>Case I</u>: $r = \frac{2\omega_o}{H_o Q}$, $\delta = 0$

$$T_{All\ Pass} = \frac{V_o'}{V_i} = k'\,\frac{S^2 - \frac{\omega_o}{Q}S + \omega_o^2}{S^2 + \frac{\omega_o}{Q}S + \omega_o^2} \tag{1.37}$$

<u>Case II</u> $r = \frac{\omega_o}{H_o Q}$, $\delta \neq 0$

$$T_{notch} = k'\,\frac{S^2 + \omega_o^2}{S^2 + \frac{\omega_o}{Q}S + \omega_o^2} \tag{1.38}$$

<u>Case III:</u> $r = \frac{\omega_o}{H_o Q}$, $\delta \neq 0$

$$T_{LP\ notch} = k \frac{S^2 + \omega_o^2(1 + k_o \delta)}{S^2 + S\frac{\omega_o}{Q} + \omega_o^2} \qquad (1.39)$$

<u>Comments</u>: The scheme implemented for obtaining all pass and notch functions is not practically feasible due to loading problems. Independent tuning of ω_o, Q and G_o is possible by trimming three resistors. The sensitivities are less than 1.

<u>Comparison with Rao's and Shawman's Circuit</u>

In Rao's circuit, the attainable Q for a specified ω_o and B is limited. In Schawman's circuit, the midband gain does not remain constant as Q is varied. For the same ω_o, and Q, the proposed circuit has lower resistance ratios. The impedance at the input of the proposed structure at low and high frequencies is approximately equal to R_1 and intermediate frequencies is slightly lower than R_1. In Schawman's circuit the input impedance except at high frequencies is affected by midband gain and is lower than the input impedance of the former for the same R_1.

1.3.5 Li and Li [14]:

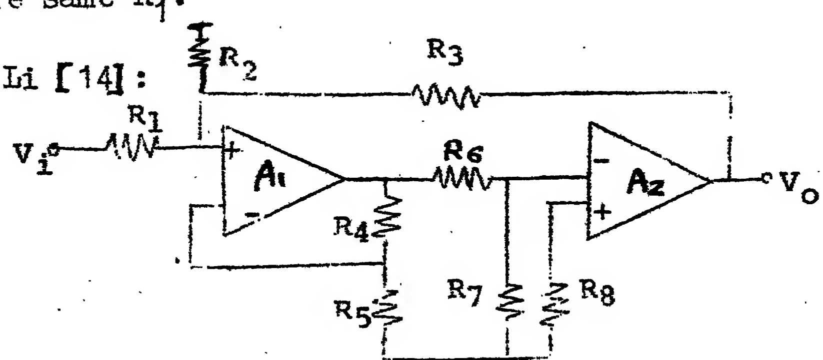


Fig. 1.12

$$T_{BP} = \frac{V_1}{V_i} = \frac{S\ R_1 R_3 B_1}{(R_1R_2+R_2R_3+R_1R_3)} / D \tag{1.40}$$

$$T_{LP} = \frac{V_o}{V_i} = \frac{R_7\ R_2\ R_3\ B_1\ B_2}{((R_6+R_7)(R_1R_2+R_2R_3+R_1R_3))} / D \tag{1.41}$$

Where

$$D = S^2 + S \frac{R_5}{R_4+R_5} B_1 + \frac{R_7\ R_1\ R_2\ B_1\ B_2}{(R_6+R_7)(R_1R_2+R_2R_3+R_1R_3)}$$

1.3.6 Soderstand [15]:

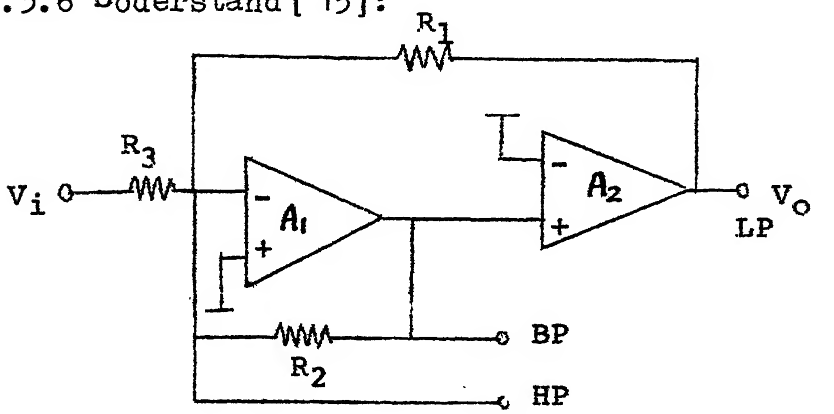


Fig.1.13

$$T_{BP} = \frac{-(B_1/R_3)S}{D} \tag{1.42}$$

$$T_{LP} = \frac{-(B_1B_2)/R_3}{D} \tag{1.43}$$

$$T_{HP} = \frac{S^2(1/R_3)}{D}$$

where

$$D = \left(\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}\right)S^2 + \left(\frac{B_1}{R_2}\right)S + \frac{B_1\ B_2}{R_1}$$

Second Order Section using Wideband Summing Amplifier

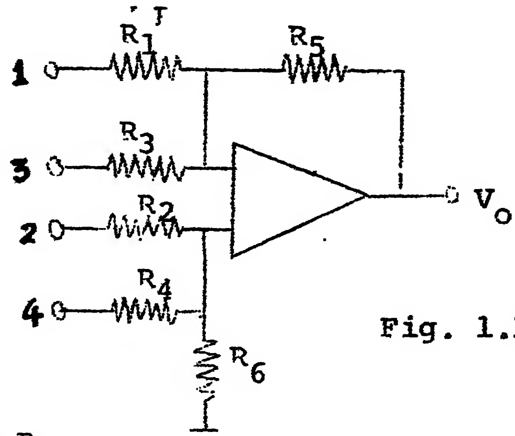


Fig. 1.14

$$K_i = \frac{\dfrac{(-1)^i\, k_i B}{k_1+k_3+1}}{s^2+\left(\dfrac{B}{k_1+k_3+1}\right)} \tag{1.45}$$

where

$$k_i = \frac{R_5}{R_i} \qquad i = 1,3$$

$$= \frac{R_5}{R_i}\left(\frac{1/R_1 + 1/R_3 + 1/R_5}{1/R_2 + 1/R_4 + 1/R_6}\right) \qquad i = 2,4$$

For wide band $\omega << \frac{B}{1+k_1+k_2}$ where $K_i \approx (-1)^i\, k_i$

General Active R Building Block using Wideband Summer

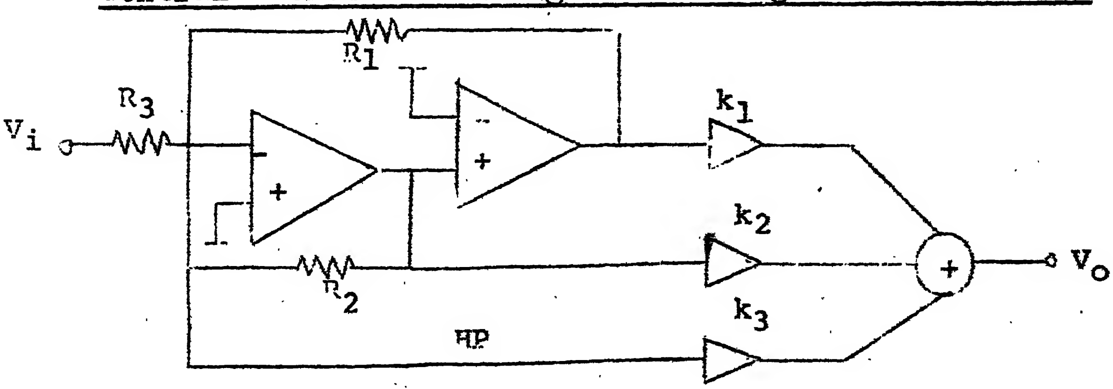
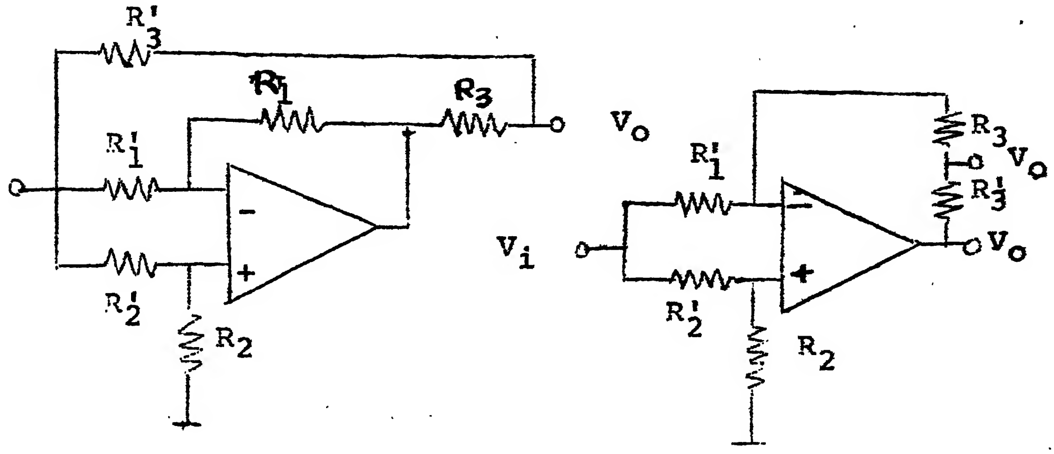


Fig.1.15

$$T_{out} = \frac{\alpha_2 S^2 + \alpha_1 S + \alpha_o}{S^2 + \frac{\omega_o}{Q} S + \omega_o^2} \tag{1.46}$$

where $\alpha_2 = k_3/R_3$, $\alpha_1 = -\frac{k_2 B_1}{R_3}$, $\alpha_o = -\frac{k_1 B_1 B_2}{R_3}$

To get B.E. function $\alpha_1 = 0$. To get third order function the summing amplifier will be a version of first order section shown in Fig. 1.16(a,b)



(a) General feedforward first order filter.

(b) General node xumming first order filter.

Fig. 1.16

$$T(s) = \frac{k_o(s + \omega_{oo})}{S + \omega_o} \qquad T(s) = \frac{k_o(s+\omega_{oo})}{S+\omega_o} \tag{1.47a,b}$$

$k_o = P_3$ $\qquad$ $k_o = P_1 (1 - P_2)$

$\omega_{oo} = 1 + P_1 + \frac{(1-P_1)(P_2-P_1)}{P_3}$ $\qquad$ $\omega_{oo} = \frac{(1-P_3)(1-P_1)P_2 + P_2(P_2-P_1)}{P_1(1-P_2)}$

$\omega_o = (1-P_1)B$ $\qquad$ $\omega_o = (1-P_1)B$

$P_1 = R_1/(R_1+R_1')$, $\quad P_2 = R_2/(R_2+R_2')$, $\quad P_3 = R_3/(R_3+R_3')$.

Comments: In choosing the Op-amp for the summing amplifier it would be desirable to choose one with the gain bandwidth product considerably larger than the amplifiers used in the basic section for the wideband case, and considerably smaller than gain bandwidth product for the third order or narrow band case.

1.3.7 Srinivasan, S. [11]:

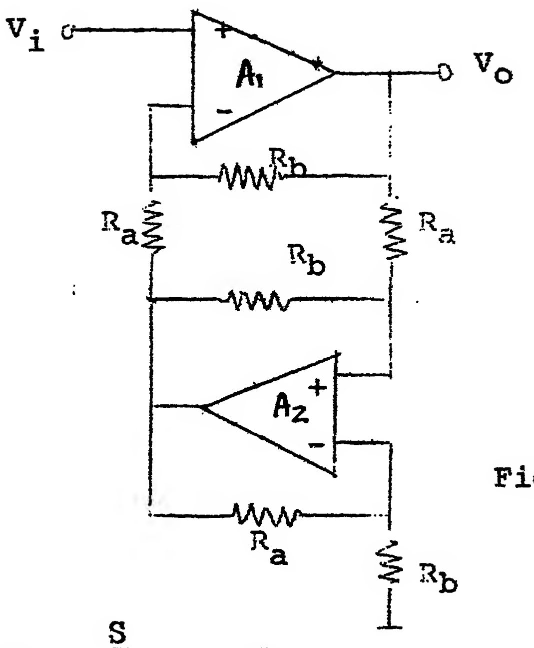


Fig. 1.17

$$T_{BP} = \frac{V_o}{V_i} = \frac{S}{S^2+S\,k + (1-k)^2} \tag{1.49}$$

For $A_o >> 2/k, >> 1/k$ and $>> k/(1-k)^2$

where $k = R_a/(R_a+R_b)$

Hence either ω_o or Q is independently chosen and fixing one fixes the other.

1.3.8 Soliman [16]:

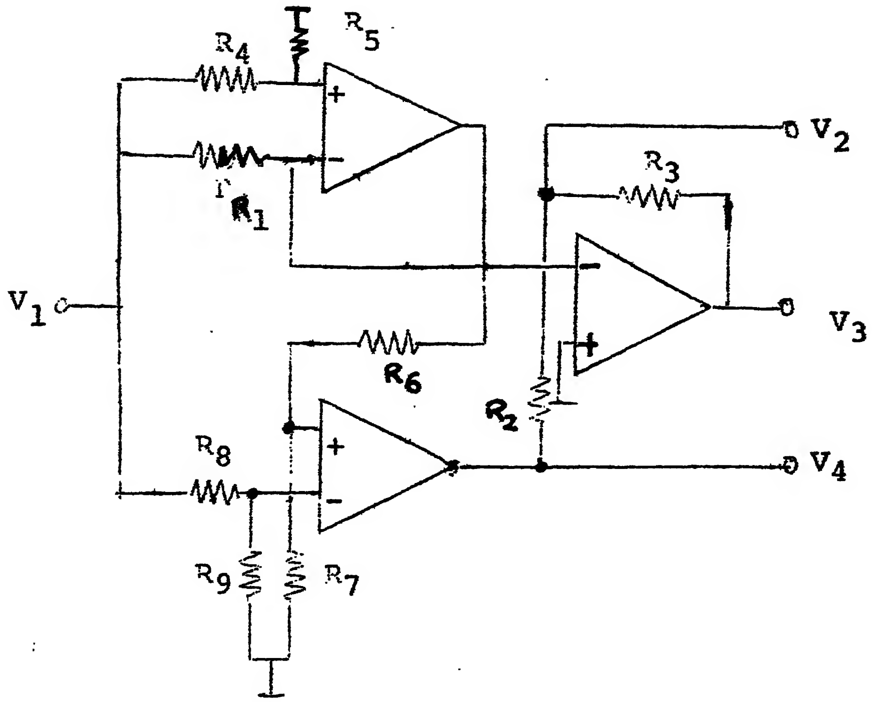


Fig.1.18

$$T_1(s) = \frac{V_2}{V_1} = k \frac{S^2 - S(\frac{\omega_z}{Q_z}) + \omega_z^2}{S^2 + S(\omega_p/Q_p) + \omega_z^2} \tag{1.50}$$

$k = a/b$, $a = R_2/R_1$, $b = 1+a + \frac{R_2}{R_3}$, $n = R_5/(R_4+R_5)$,

$m = R_7/(R_6+R_7)$, $P = R_9/(R_8/R_9)$.

$\omega_z^2 = m\ n\ B_1\ B_2\ \frac{1}{a}$, $\qquad Q_z = \frac{1}{P}(m\ n\ a\ \frac{B_1}{B_2})^{\frac{1}{2}}$

$\omega_p^2 = m\ B_1\ B_2\ \frac{1}{b}$, $\qquad Q_p = ((m\ b\ B_1B_2)^{1/2}/(b-a-1)B_3)^{\frac{1}{2}}$

Case I: All pass transfer function

$n = k$

$$P = \frac{B_3}{B_2} \frac{a(b-a-1)}{b}$$

Case II: General notch filter

$P = 0$

$n = k$ notch filter

$n > k$ lowpass notch filter

$n < k$ highpass notch filter.

Case III: Highpass filter

$P = 0$

$n = 0$

Case IV: Bandpass filter

$P = 0, \quad n = 0$

$$\frac{V_3}{V_1} = \frac{-H_1 S}{S^2 + S(\frac{\omega p}{Q_p}) + {}^2_p} \tag{1.51}$$

where $H_1 = k B_3$.

$$|G_o| = \frac{a}{b-a-1} = \frac{R_3}{R_1}$$

Case V: Low-pass filter

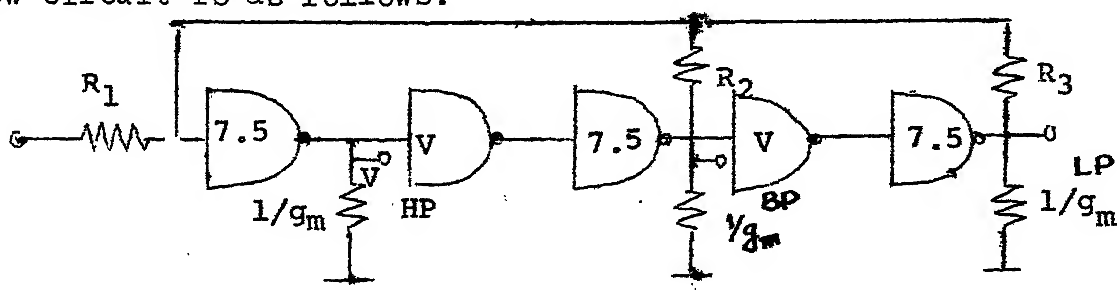
$$\frac{V_4}{V_1} = \frac{-H_z}{S^2 + S\frac{\omega p}{Q_p} + w_p^2}, \quad H_z = mkB_1B_2 \tag{1.52}$$

$|G_o| = a = R_2/R_1$.

1.4 CMOS Active R Filters 17

Active R filters generally are simpler to design and may be operated at higher frequencies than standard active RC filters. However, in active R filters loading and stability problems are severe and also often required special tuning procedure. CMOS transistor arrays eliminate loading and stability problems.

Soderstand has extended his previous work 15 by replacing Op-amps by CMOS transistor pair and has used buffer to obtain inverting gain as well as to avoid loading. The new circuit is as follows:



The buffers are connected to ± 7.5 constant supply while amplifiers are connected to variable supply for voltage tuning of filter.

CMOS active R filter technique results in circuits which could find practical applications in relatively low Q application whose voltage tunability relatively low frequency response or CMOS techniques are desirable features.

CHAPTER 2

OPERATIONAL AMPLIFIER MODELLING

2.1 Introduction

Frequently active filters are designed by treating the active element as ideal. It is assumed that the active element used in the realisation is an Op-amp such as μA741. Ideally the Op-amp is an infinite gain dc amplifier with infinite input impedance, zero output impedance and in general, differential input and differential output capabilities. In practice the Op-amp can be realised with a dc gain of 10^4 or more, the input impedance greater than 100 K and output impedance less than 100 Ω. Moreover, the open loop gain A must be limited to a one pole frequency characteristics in order to obtain stability when feedback is applied around the Op-amp, when they are used as inverting and non-inverting amplifier.

Neglecting this frequency dependence of the amplifier gain in the design of active filters can lead to serious accuracy, sensitivity and stability problems which usually restrict the filter performances to low Q and low frequency operation. On the other hand, the amplifier poles themselves can be utilised in realising second order bandpass filter which reduces instability problems as well as makes the circuit suitable for integration by reducing external capacitor. The idea of utilising the gain roll off of the amplifier as a

as a capacitor can be extended to completely eliminate capacitors in the active filter design.

In order to ensure stable operation in a closed loop application, operational amplifiers are designed to have a frequency response such as shown in Fig.2.1. This is often accomplished by using either an internally compensated Op-amp (μA741) or through an external compensating network (μA709).

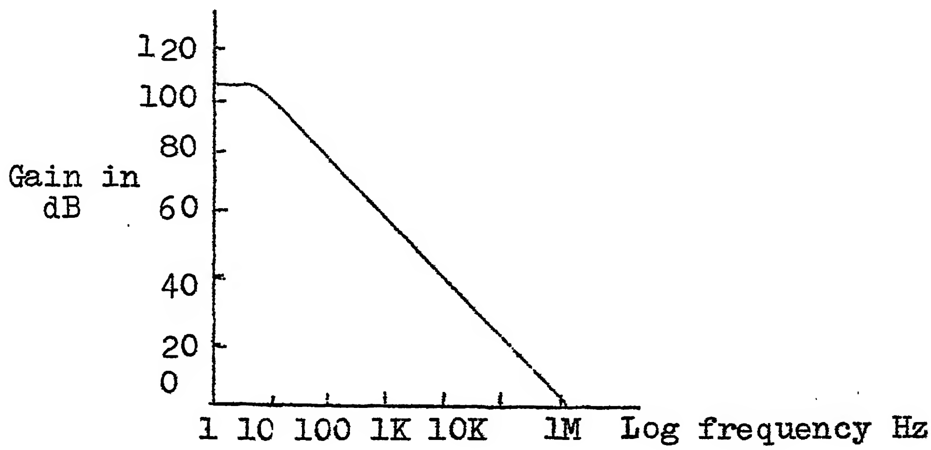


Fig.2.1 Open loop gain response of μA741.

In this thesis, we assume the one pole model of the Op-amp, characterised by

$$A(s) = \frac{A_o}{1 + s\tau} = \frac{B}{s + \frac{1}{\tau}} \tag{2.1}$$

2.2 Open Loop Gain Measurement

In this section two methods of measuring open loop dc gain A_o and the pole frequency $1/\tau$ are discussed.

Method-1

Consider the circuit shown in Fig.2.2. The input admittance is given by

$$Y_i = \frac{1}{Z_i} = \frac{1}{RA_o} + \frac{s\tau}{RA_o} \qquad (2.2)$$

assuming $A_o >> 1$ and $A_o > |s\tau|$. Here R is chosen small so that measurements could be made accurately. The real and imaginary parts of the input admittance are measured from which A_o and τ can be calculated.

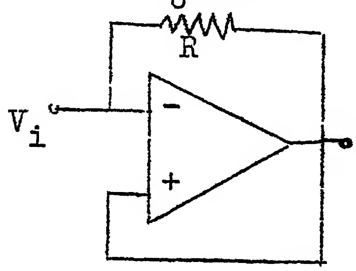


Fig.2.2

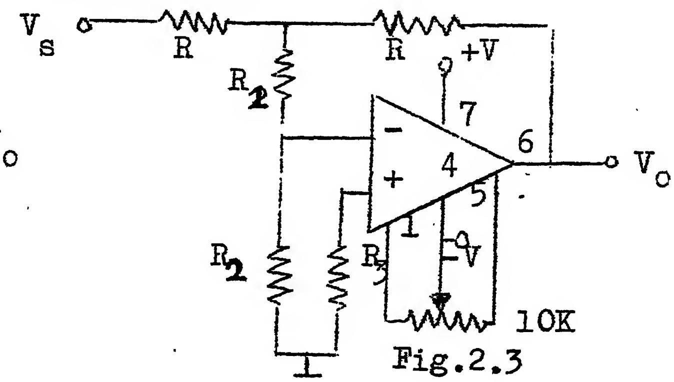


Fig.2.3

In this thesis the method-2 outlined below is used to find A_o and τ of the Op-amps.

Method-2

This method uses the circuit given in Bur-Brown [18] to find out A_o and τ of the Op-amp. In this method the open loop gain of the Op-amp is plotted against the frequency and the curve is fitted to the equation

$$A = \frac{A_o}{1+s\tau}$$

The circuit diagram is shown in Fig.2.3. This circuit uses closed loop condition so that V_i and V_o are of same level.

The equation connecting V_o and V_r is obtained as follows:

$$V_o = -V_i\, A$$
$$V_i = V_r \frac{R_2}{R_1+R_2} \qquad (2.3)$$
$$V_o = -V_r\, A \frac{R_2}{R_1+R_2}$$

From (2.3) an expression is obtained for the open loop gain A as

$$A = -\frac{V_o}{V_r} \frac{R_1+R_2}{R_2} \qquad (2.4)$$

Using (2.4) open loop gain A is plotted against frequency by varying the input frequency and measuring V_o and V_r for each frequency.

In this circuit open loop voltage gain is measured under closed loop conditions with a specified output level by measuring the voltage V_r associated with the small input signal voltage V_i of the amplifier. The input offset voltage without feedback, causes the output to go to saturation. This could be eliminated by either giving the required bias voltage at the non-inverting input of the Op-amp, or by having negative feedback. In this circuit negative feedback is used to suppress the effect of input offset voltage. A voltage divider is used so that the voltage V_r is not very small and thus measurable accurately.

The order of output voltage will be the same as V_s because of feedback. The input voltage V_i will be in V at low frequencies and the voltage V_r will be proportional to V_i the proportionality constant being $\frac{R_1+R_2}{R_2}$.

Open loop dc gain A_o and first pole angular frequency $1/\tau$ determine the resistors used in the circuit proposed and thereby determine G,Q and ω_o in the bandpass experiment. Therefore to get predictable experimental results A_o and should be measured accurately. Moreover, $T = \tau/A_o$ comes in the denominator of the expression for centre frequency, ω_o, of the bandpass filter. Thus if A_o and τ are not measured accurately ω_o will change from the design value. These factors constraints us to measure A_o and τ as accurately as possible. Hence the measurement of voltage was done using VTVM. The output voltage V_o and voltage V_r will be of different order and so the VTVM calibrations were checked before proceeding with the measurements. This was done as follows.

A constant dc voltage, say 1V, was supplied to the millivoltmeter in the 10V range and measured. Then the range was changed to 3V scale and the voltage was again measured. This was repeated with all the lower scales till 1 mV range and the consistency of calibration was proved.

According to the model, Op-amp should have only one pole before the gain bandwidth product or 0 dB frequency.

However, it is sufficient if the Op-amp has only one pole before 10 times the operating frequency and the roll off of 6 dB/octave. This also means that the phase response should be -45°(lag) at the pole frequency and attain -90°(lag) at the final frequency. As this only confirms the model this measurement was made using oscilloscope. Since the pole frequency should be measured as accurately as possible, frequency was measured using digital frequency meter.

As the frequency is increased voltage V_r is increased since the gain falls. In order to keep the output in the same order throughout the experiment the voltage V_i has to be increased. This is done by increasing the ratio of resistors R_2/R_1. R_3 is chosen as equal to R_2 to reduce the offset voltage.

2.3 Expreimental Procedure

To start with the experiment, the input was grounded and the output offset was adjusted to zero voltage by using the offset null arrangement between terminal 1 and 5 as shown in Fig.2.3.

A small dc positive voltage of the order of volts was applied at the input. The voltages at V_o and V_r were measured and the gain was found out. Since a positive voltage was being applied at the inverting input, the output should had been negative. But no polarity inversion was found.

Then a small negative voltage of the order of volts was applied at the input and gain was calculated. The negative voltage was increased step by step and the output was found to be increasing positively. This confirmed the polarity inversion. Maintaining the linearity, the gain was found out at different input voltages. Thus the dc gain of the Op-amp was found out.

Then the low frequency gain and phase was measured. To start with the experiment a sinusoidal signal generator was connected to the input and voltage at V_o and V_r were measured using VTVM. It was found that at low frequency the output voltage V_o was always lagging the input voltage V_r by 270°. This inconsistancy in phase and gain response was removed by the following way. By the offset arrangement, a small positive voltage of the order of 2V was made to stand at the output with the input grounded. Then the low frequency signal was given at the input. It was found that the output was lagging the input by 180° at 0.1 Hz and below that. Now as the frequency was increased to 1 Hz and beyond that, the output phase lag increased beyond 180° and gradually became 270° beyond 200 Hz till 100 KHz. The first pole was confirmed to be in between 1 Hz and 10 Hz, where the phase lag became 235°.

The gain and phase response was measured till 1 MHz. The output voltage was maintained in the order of volts.

When the voltage V_r became sufficiently high, R_1 and R_2 were changed and the experiment repeated. In this way the frequency response of two Op-amps were measured. It was found that the single pole model was obeyed, as seen from the gain response plot shown in graph 1, till the 0 dB frequency. But the phase response did not confirm with the one pole model beyond 100 KHz, as the phase lag was found to be increasing more than 270°, where it should had stayed till the 0 dB frequency. This is due to the second pole effect. The phase response was measured till 1.5 MHz and the second pole was found out by fitting the phase response bringing in the effect of second pole.

This excess phase shift limits the performance of μA741 to 100 KHz in filter circuits. The second pole model was to be considered when operating in the range between 100 KHz and 1 MHz. Since the second pole has negligible effect on gain response while effecting the phase response drastically, a second pole mode of A741 may be thought of as follows.

$$A(s) = \frac{A_o\,(1 - S/2\omega_2)}{(1+s\,\tau)\,(1 + S/2\omega_2)} \qquad (2.5)$$

where

A_o = open loop dc gain

$1/\tau$ = open loop pole frequency radians/sec.

ω_2 = second pole frequency radians/sec.

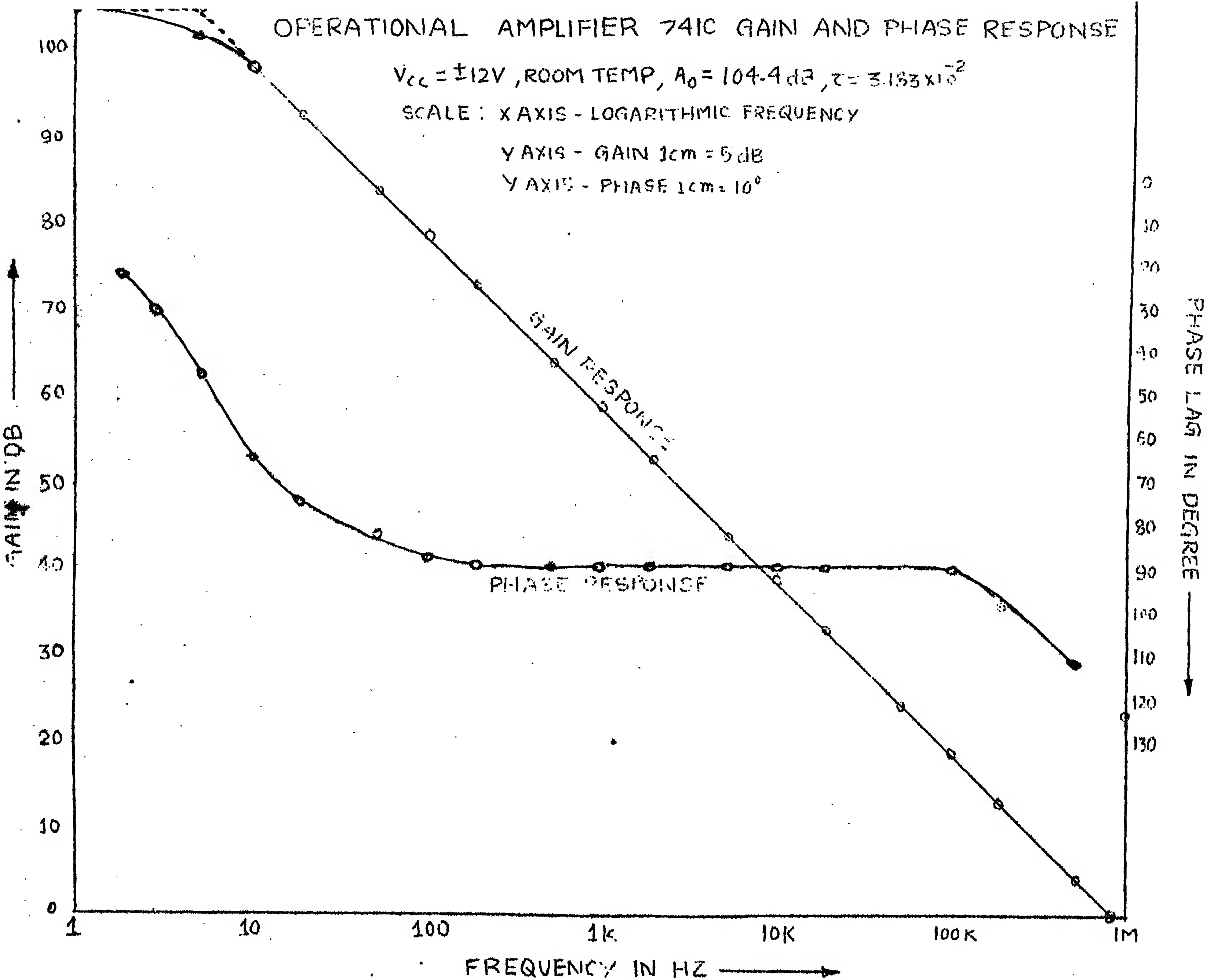
ω = operating frequency radians/sec.

Three Op-amps were taken for modelling. The open loop gain and first pole found by experiments are given below.

$$A_{o1} = 166.6\times10^{3} \qquad A_{o2} = 166.6\times10^{3} \qquad A_{o3} = 180\times10^{3}$$

$$\tau_1 = 3.183\times10^{-2}\text{ secs} \quad \tau_2 = 3.183\times10^{-2}\text{ secs} \quad \tau_3 = 3.183\times10^{-2}\text{ secs}$$

Two Op-amps were found to be matched. Substituting the values of A_o and τ in equation (2.1), it is found that the theoretical and experimental open loop gain and phase response match very closely, as given in Tables 3.1 and 3.2.



OPERATIONAL AMPLIFIER 741C GAIN AND PHASE RESPONSE
Vcc = ±12V, ROOM TEMP, A0 = 104.4 dB, τ = 3.183x10^-2
SCALE : X AXIS - LOGARITHMIC FREQUENCY
Y AXIS - GAIN 1cm = 5dB
Y AXIS - PHASE 1cm = 10°
GAIN RESPONSE
PHASE RESPONSE
GAIN IN DB
PHASE LAG IN DEGREE
FREQUENCY IN HZ
100
90
80
70
60
50
40
30
20
10
0
1
10
100
1k
10K
100K
1M
0
10
20
30
40
50
60
70
80
90
100
110
120
130

TABLE 2.1

FREQUENCY RESPONSE OF OP-AMP μA741C (1,2)

Values of R_1 & R_2	Freq. Hz	V_r mV	V_o mV	A $=(V_o/V_i)$ x 10^3	A dB experimental	A dB Theoretical
	0	-60.00	+10	166.6	104.4	104.4
	0	-47.50	+ 8	168.4	104.5	104.4
	0	-30.0	+ 5	166.0	104.4	104.4
R_1=100Kohm	5	8.47	+ 1	118.0	101.43	101.39
R_2=100ohm	6	10.3	1	97.3	99.76	100.52
	7	10.7	1	93.2	99.38	99.68
	8	12.2	1	82.0	98.27	98.87
	10	13.2	1	75.75	97.58	97.41
	20	24.0	1	41.66	92.39	92.09
	50	58.0	1	17.24	84.73	84.35
	100	1.29	1	8.527	78.6	78.36
	200	2.6	1	4.23	72.53	72.34
	500	6.3	1	1.746	64.8	64.4
	1K	12.8	1	0.8593	58.68	58.37
	2K	25.6	1	0.429	52.66	52.34
	5K	65.0	1	0.169	44.57	44.4
	10K	132.0	1	0.0833	38.41	38.38
	20K	7.9	0.03	41.77	32.4	32.36
R_1=10K	50K	19.8	0.03	16.66	24.4	24.4
R_2=1K	100K	39.0	0.03	8.46	18.55	18.38
		(V-) Inverting terminal mV)				
	200K	7.0	0.03	4.285	12.64	12.36
	400K	14.0	0.03	2.143	6.62	6.34
	600K	21.1	0.03	1.422	3.05	2.81
	800K	28.1	0.03	1.06	0.506	0.32
	900K	31.5	0.03	0.952	-0.427	-0.703
	1M	34.9	0.03	0.86	-1.31	-1.6

TABLE 2.2

PHASE RESPONSE OF OP-AMP μA741C (1,2)

f Hz	Phase angle experimental lag (°)	Phase angle theoretical lag (°)
0	0	0
2	21.6	21.8
3	30.6	30.96
5	45.0	45.0
6	50.4	50.19
7	54.0	54.46
8	61.2	57.99
10	64.8	63.4
20	75.6	75.96
50	82.8	84.29
100	88.0	87.13
200	90.0	88.56
500	90.0	89.42
1K	90.0	90.0
⋮	⋮	⋮
100K	90.0	90.0

Continued...

(Table 2.2 continued)

f Hz	Phase angle experimental lag (°)	Phase angle theoretical lag (°)	Phase angle due to 2nd pole at 1.5MHz(lag°)
200K	99.0	90.0	98.13
300K	104.03	90.0	102.09
400K	108.43	90.0	106.0
500K	111.60	90.0	109.65
600K	112.5	90.0	113.2
700K	116.5	90.0	116.56
800K	120.0	90.[illegible]	119.74
1 M	124.0	90.0	125.53
1.1M	126.0	90.0	128.15
1.2M	135.0	90.0	130.6
1.3M	141.0	90.0	132.8
1.4M	145.0	90.0	135.0
1.5M	147.0	90.0	136.9

TABLE 2.3

FREQUENCY RESPONSE OF OP-AMP μA741C(3)

Values of R_1 & R_2	Freq. Hz	V_r mV	V_o V	A $=(V_o/V_i) \times 10^3$	A dB Experimental	A dB Theoretical
	0.0	-50.0	+9	180.0	105.1	105.1
	0.0	-33.3	+6	180.18	105.114	105.1
	0.0	-19.4	+3.5	180.4	105.124	105.1
	5.0	7.66	1	130.55	102.31	102.09
R_1=100K-ohm	6.0	8.9	1	112.36	101.01	101.23
R_2=100 ohm	7.0	9.73	1	102.77	100.23	100.39
	8.0	10.75	1	93.02	99.37	99.59
	10.0	12.40	1	80.64	98.13	98.11
	20.0	23.60	1	42.37	92.54	92.80
	50.0	54.50	1	18.35	85.27	85.06
	100	1.225	1	8.98	79.06	79.07
	200	2.45	1	4.49	73.05	73.06
	500	6.1	1	1.8	65.1	65.1
	1K	12.2	1	0.901	59.09	59.08
	2K	24.4	1	0.4508	53.08	53.06
	5K	61.0	1	0.1803	45.12	45.1
	10K	122.0	1	0.9	39.1	39.08
	20K	7.32	0.03	45.08	33.07	33.06
	50K	18.34	0.03	17.99	25.1	25.1
	100K	36.6	0.03	9.016	19.1	19.08
		(V-)Inverting terminal mV)				
	200K	6.7	0.03	4.497	13.05	13.06
	500K	16.7	0.03	1.796	5.08	5.1
	600K	20.0	0.3	1.5	3.52	3.52
	800K	26.7	0.03	1.123	1.007	1.023
	900K	30.0	0.03	1.0	0.0	0.0

TABLE 2.4

PHASE RESPONSE OF OP-AMP μA741C(3)

f Hz	Phase angle experimental lag (°)	Phase angle theoretical lag (°)
0	0	0
2	21.5	21.8
3	30.8	30.96
5	45.0	45.0
6	50.6	50.19
7	53.8	54.46
8	59.2	57.99
10	64.8	63.4
20	75.8	75.96
50	82.8	84.29
100	88.0	87.13
200	90.0	88.56
500	90,0	89.42
1K	90.0	90.0
.	.	.
.	.	.
.	.	.
100K	90.[illegible]	90.0

CHAPTER 3

DEVELOPMENT OF PROPOSED CIRCUIT MODEL

The basic circuit model as shown in Fig.3.1 was realised by Caparelli [3] to obtain bandpass transfer function. Block 1 and Block 2 realise lowpass function. Block 3 is a constant gain amplifier. Block 2 in combination with Block 3 realise the highpass function. Then cascading lowpass and highpass blocks, we get bandpass function. Positive feedback is applied to get high Q.

3.1 Basic Circuit Model

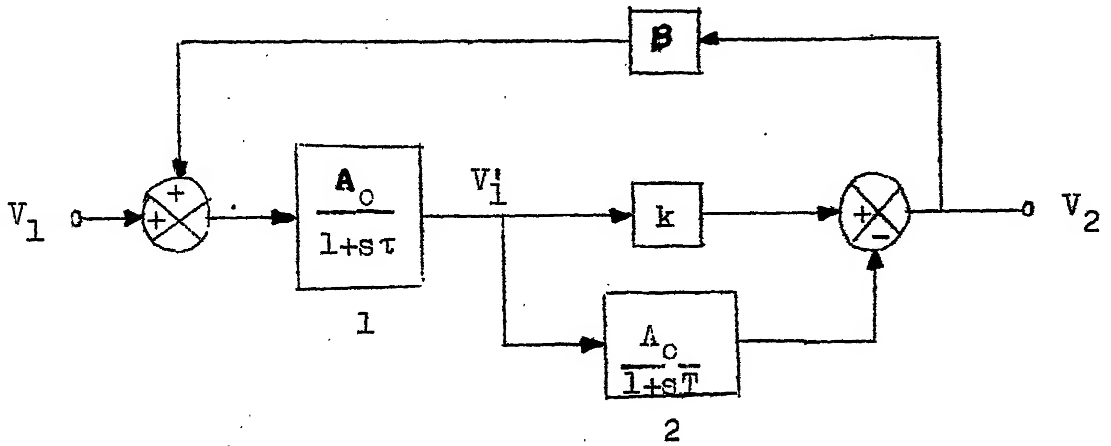


Fig.3.1

Transfer Function

Without 'β' feedback:

$$V_1' = V_1 \frac{A_o}{1+s\tau} \tag{3.1}$$

$$V_2 = V_1'(k - \frac{A_o}{1+s\tau}) \tag{3.2}$$

Substituting (3.1) in (3.2) and assuming $k = A_o$,

$$V_2 = V_1 \frac{A_o^2 \tau s}{(s\tau+1)^2} \tag{3.3}$$

$$H(s) = \frac{V_2}{V_1} = \frac{A_o^2 \tau s}{(s \quad +1)^2} \tag{3.4}$$

With 'β' feedback:

$$T(s) = \frac{H(s)}{1-H(s)\beta} = \frac{A_o^2 \tau s}{(s \; +1)^2 - A_o^2 \tau s \beta}$$

$$= \frac{A_o^2 \tau s}{s^2 \tau^2 + s\tau(2-A_o^2 \beta)+1}$$

$$= \frac{(A_o^2/\tau)s}{s^2 + \frac{s}{\tau}(2-A_o^2 \beta) + \frac{1}{\tau^2}} \tag{3.5}$$

$$\omega_o = \frac{1}{\tau} \tag{3.6}$$

$$Q = \frac{1}{2-A_o^2 \beta} \tag{3.7}$$

$$G_o = \frac{A_o^2}{2 - A_o^2 \beta} \tag{3.8}$$

If $\beta = 2/A_o^2$, Q can be made infinite. However, ω_o is fixed by the pole frequency of the amplifier. To get variable ω_o the circuit realisation of the basic building block $A_o/(s\tau+1)$ can be made as follows.

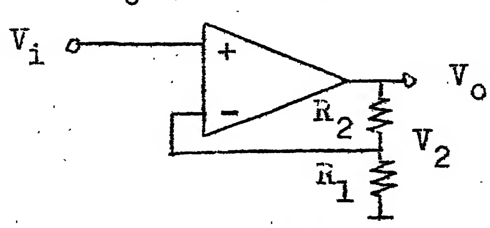


Fig.3.2

<u>Transfer function</u>

$$V_2 = V_o \frac{R_1}{R_1+R_2} \tag{3.9}$$

$$V_i - V_2 = \frac{V_o}{\frac{A_o}{s\ +1}} \tag{3.10}$$

$$V_i = V_o\left(\frac{R_1}{R_1+R_2} + \frac{s\tau+1}{A_o}\right)$$

$$\frac{V_o}{V_i} = \frac{1}{\frac{R_1}{R_1+R_2} + \frac{s\tau}{A_o} + \frac{1}{A_o}} \approx \frac{1}{\frac{s\tau}{A_o} + \frac{R_1}{R_1+R_2}}$$

$$\left(\text{assuming } A_o \gg \frac{R_1+R_2}{R_1}\right)$$

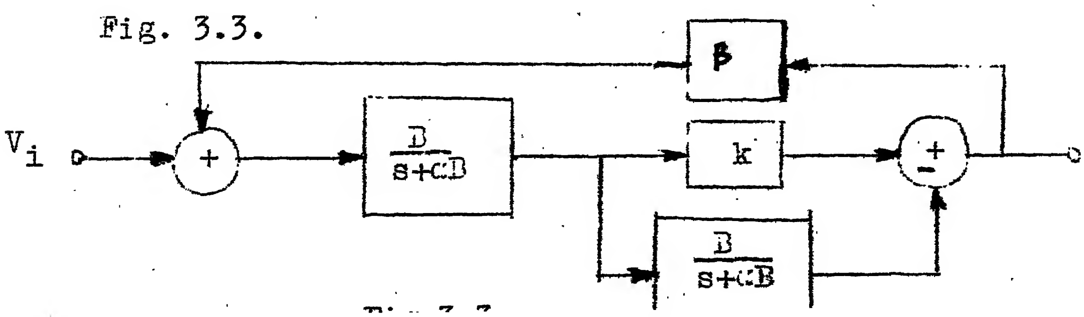
$$\frac{V_o}{V_i} = \frac{A_o/\tau}{s + \frac{R_1}{R_1+R_2}\,\frac{A_o}{\tau}} = \frac{B}{s+\alpha B} \tag{3.11}$$

where $\alpha = \frac{R_1}{R_1+R_2}$

So by varying the resistor ratio R_1/R_2, the pole frequency can be changed as desired.

3.2 New Circuit Model

The new circuit model bringing in the provision of electronic tuning in the basic lowpass block is shown in Fig. 3.3.



<u>Transfer function</u>

Without 'β' feedback:

$$H(s) = \frac{V_2}{V_i} = \frac{B}{s+\alpha B}\left(k - \frac{B}{s+\alpha B}\right)$$

$$\left(\text{assuming } k = \frac{1}{\alpha} = \frac{R_1+R_2}{R_1}\right)$$

$$H(s) = \frac{V_2}{V_i} = \frac{s\,\frac{B}{\alpha}}{(s+\alpha B)^2} \qquad (3.12)$$

With 'β' feedback:

$$T(s) = \frac{H(s)}{1 - H(s)\beta} = \frac{s\,\frac{B}{\alpha}}{(s+\alpha B)^2 - s\beta\,\frac{B}{\alpha}}$$

$$= \frac{s\,\frac{B}{\alpha}}{s^2 + s\,B\left(2\alpha - \frac{\beta}{\alpha}\right) + (\alpha B)^2} \qquad (3.13)$$

$$\omega_o = \alpha B \qquad (3.14)$$

$$Q = \frac{1}{\alpha\text{[illegible]} - \frac{\beta}{\alpha 2}} \qquad (3.15)$$

$$G_o = \frac{1}{\alpha^2\left(2 - \frac{\beta}{\alpha^2}\right)} \qquad (3.16)$$

From the above expression we see that ω_o and Q can be changed independently by changing α and β respectively. So the above scheme seems suitable. But there are certain disadvantages as follows, with the solution given.

Since we have assumed $K = \frac{1}{\alpha} = \frac{R_1+R_2}{R_1}$, a constant gain amplifier over the entire range of frequency of interest is required. So the scheme is modified as shown in Fig.3.4.

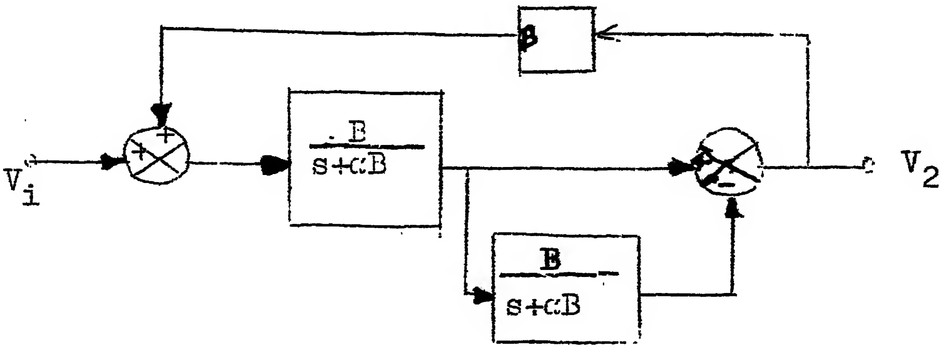


Fig.3.4

$$T(s) = \frac{sB}{s^2+sB(2\alpha-\beta)+(\alpha B)^2} \tag{3.17}$$

3.3 Three Op-amp Circuit

From the above scheme, we see that four Op-amps are required to carryout the different functions indicated. Minimisation of passive and active components, is a desirable feature in the realisation of active filters. If we can combine the functions of 1 and 2 and consider 4 as a flat summer, a circuit with three Opamps can be realised as follows.

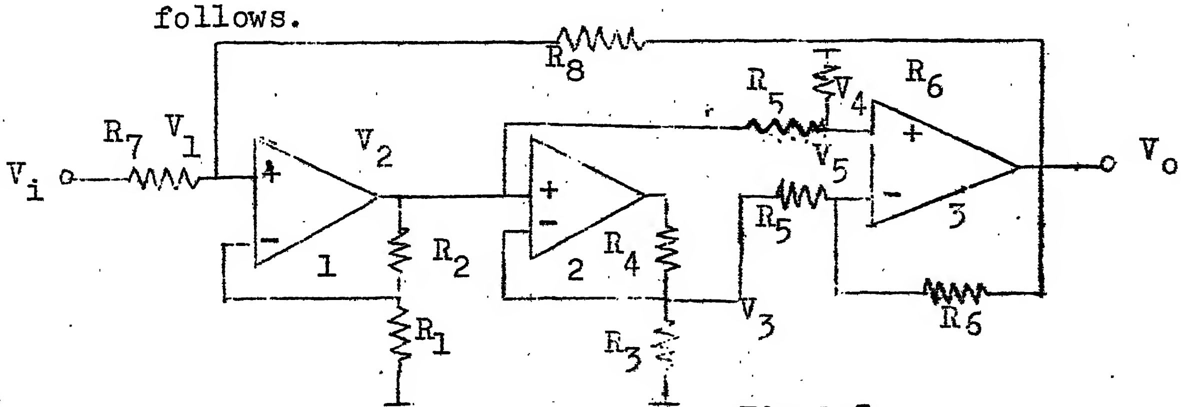


Fig.3.5

3.3.1 Analysis considering a flat summer

Let $\alpha = \frac{R_1}{R_1+R_2}$, $\beta = \frac{R_3}{R_3+R_4}$, $\gamma = \frac{R_5}{R_5+R_6}$, $\delta = \frac{R_7}{R_7+R_8}$

$$V_1 = (V_i(1-\delta) + V_o\,\delta \tag{3.18}$$

$$V_1 - \alpha V_2 = V_2\left(\frac{1+s\,\tau_\alpha}{A_{o\alpha}}\right)$$

where $A_{o\alpha}$ = open loop gain of Op-amp 1

τ_α = open loop pole frequency of Op-amp 1

$$V_2 = V_1 \frac{A_{o\alpha}/\tau_\alpha}{s+\alpha\,(A_{o\alpha}/\tau_\alpha)} = V_1 \frac{B_\alpha}{s+\alpha B_\alpha} \tag{3.19}$$

where B_α = gain bandwidth product of Op-amp 1.

$$V_2 - V_3 = \frac{V_3}{\beta}\left(\frac{1+s\,\tau_\beta}{A_{o\beta}}\right)$$

$$V_2 = V_3\left(1+\frac{s\,\tau_\beta}{\beta A_{o\beta}}\right) \quad \text{for } \beta A_{o\beta} \gg 1$$

$$V_3 = V_2 \frac{\beta B_\beta}{s+\beta B_\beta} \tag{3.20}$$

$$V_o = (V_2 - V_3)\frac{R_6}{R_5} = (V_2 - V_3)\left(\frac{1-\gamma}{\gamma}\right) \tag{3.21}$$

Substituting (3.18), (3.19) and (3.20) in (3.21)

$$V_o = V_2 \frac{s}{s+\beta B_\beta} \frac{1-\gamma}{\gamma}$$

$$= V_1 \frac{B_\alpha}{s+B_\alpha} \frac{s}{s+\beta B_\beta} \frac{1-\gamma}{\gamma}$$

$$= [V_i(1-\delta) + V_o\delta] \frac{B_\alpha}{s+\alpha B_\alpha} \frac{s}{s+\beta B_\beta} \frac{1-\gamma}{\gamma}$$

$$V_o[1-\delta(\frac{1-\gamma}{\gamma})\frac{B_\alpha}{s+\alpha B_\alpha}\ \frac{s}{s+\beta B_\beta}] = V_i(1-\delta)(\frac{1-\gamma}{\gamma})\frac{B_\alpha}{(s+\alpha B_\alpha)}\ \frac{s}{(s+\beta B_\beta)}$$

$$\frac{V_o}{V_i} = \frac{(1-\delta)(\frac{1-\gamma}{\gamma})B_\alpha\ s}{(s+\alpha B_\alpha)(s+\beta B_\beta) - \delta(\frac{1-\gamma}{\gamma})B_\alpha s}$$

$$= \frac{(1-\delta)(\frac{1-\gamma}{\gamma})B_\alpha\ s}{s^2+s[\alpha B_\alpha+\beta B_\beta-\delta(\frac{1-\gamma}{\gamma})B_\alpha] + \alpha\beta B_\alpha B_\beta} \tag{3.22}$$

$$\omega_o = \sqrt{(\alpha\beta B_\alpha B_\beta)} \tag{3.23}$$

$$Q = \frac{\omega_o}{\alpha B_\alpha + \beta B_\beta - \delta(\frac{1-\gamma}{\gamma})B_\alpha} \tag{3.24}$$

$$G_o = (1-\delta)(\frac{1-\gamma}{\gamma})B_\alpha\ \frac{Q}{\omega_o} \tag{3.25}$$

3.3.2 Analysis considering one pole model of summer

$$V_o = (V_4 - V_5)\ \frac{A_{o\gamma}}{1+s\tau_\gamma} \tag{3.26}$$

where $A_{o\gamma}$ = open loop gain of summer.

$1/\tau_\gamma$ = open loop pole frequency of summer

$$V_o = [V_2(1-\gamma) - V_3(1-\gamma) - V_o\gamma]\ \frac{A_{o\gamma}}{1+s\tau_\gamma}$$

$$= (V_2-V_3)\ \frac{B_\gamma}{s+\gamma B_\gamma}(1-\gamma) \quad \text{for } A_{o\gamma} >> \frac{1}{\gamma},\ |s| >> \frac{1}{\tau_\gamma}$$

$$= V_2\ \frac{s}{s+\beta B_\beta}\ \frac{B_\gamma}{s+\gamma B_\gamma}(1-\gamma)$$

$$= V_1\ \frac{B_\alpha}{s+\alpha B_\alpha}\ \frac{s}{s+\beta B_\beta}\ \frac{B_\gamma}{s+\gamma B_\gamma}(1-\gamma)$$

$$= [V_i(1-\delta)+V_o\delta]\ \frac{B_\alpha}{s+\alpha B_\alpha}\ \frac{s}{s+\beta B_\beta}\ \frac{B_\gamma}{s+\gamma B_\gamma}(1-\gamma)$$

$$\frac{V_o}{V_i} = \frac{s(1-\delta)(1-\gamma)B_\alpha B_\gamma}{(s+\alpha B_\alpha)(s+\beta B_\beta)(s+\gamma B_\gamma) - s\delta(1-\gamma)B_\alpha B_\gamma}$$

$$= \frac{s(1-\delta)(1-\gamma)B_\alpha B_\gamma}{s^3+s^2(\alpha B_\alpha+\beta B_\beta+\gamma B_\gamma)+s[\gamma B_\gamma(\alpha B_\alpha+\beta B_\beta)+\alpha\beta B_\alpha B_\beta-\delta(1-\gamma)B_\alpha B_\gamma] + \alpha\beta\gamma B_\alpha B_\beta B_\gamma} \tag{3.27}$$

3.3.3 Error calculation

Let $\gamma = \frac{1}{2}$ (unity gain summer),

$B_\alpha = B_\beta = B$, $\alpha B_\alpha = \beta B_\beta = \omega_o$

Ideal expression

$$= \frac{s(1-\delta)B}{s^2+s\alpha B[2-\frac{\delta}{\alpha}]+(\alpha B)^2} .$$

Normalising $s/\alpha B = s/\omega_o = s_n$

Ideal expression

$$= \frac{s_n(1-\delta)/\alpha}{s_n^2 + \frac{s_n}{Q} + 1} \quad \text{where } Q = 1/(2-\frac{\delta}{\alpha}) \tag{3.28}$$

Non Ideal expression

$$= \frac{s(1-\gamma)(1-\delta)BB_\gamma}{s^3+s^2\alpha B[2+\frac{\gamma B_\gamma}{\alpha B}]+s\gamma B_\gamma \alpha B[2+\frac{\alpha B}{\gamma B_\gamma}-(\frac{1-\gamma}{\gamma})\frac{\delta}{\alpha}]+\alpha^2B^2\gamma B_\gamma}$$

(Dividing the numerator and denominator by γB_γ)

$$= \frac{s(\frac{1-\gamma}{\gamma})(1-\delta)B}{\frac{1}{\gamma B_\gamma}s^3+s^2[1+2\frac{\alpha B}{\gamma B_\gamma}]+s\alpha B[2+\frac{\alpha B}{\gamma B_\gamma}-(\frac{1-\gamma}{\gamma})\frac{\delta}{\alpha}] + \alpha^2B^2}$$

(After normalisation)

$$= \frac{s_n(1-\delta)/\alpha}{s_n^3(\frac{\alpha B}{\gamma B_\gamma})+s_n^2[1+\frac{2\alpha B}{\gamma B_\gamma}]+s_n[\frac{1}{Q}+\frac{\alpha B}{\gamma B_\gamma}]+1} \tag{3.29}$$

Let $\frac{\alpha B}{\gamma B_\gamma} = \frac{1}{Qx}$ or $F_\gamma = \frac{\alpha BQx}{\gamma}$, eqn.(3.29) reduces to

Non-Ideal expression

$$= \frac{s_n(\frac{1-\delta}{\alpha})}{s_n^3 \frac{1}{Qx} + s_n^2[1+ \frac{2}{Qx}] + s_n[\frac{1}{Qx}+ \frac{1}{Q}] + 1} \qquad (3.30)$$

Dividing (3.30) by (3.28)

$$\frac{\text{Non-Ideal}}{\text{Ideal}} = \frac{s_n^2 + s_n \frac{1}{Q} + 1}{s_n^3 \frac{1}{Qx} + s_n^2[1+ \frac{2}{Qx}] + s_n \frac{1}{Q}[1+ \frac{1}{x}] +1} \qquad (3.31)$$

From the expression (3.31), comparing the numerator and denominator, we see that to achieve the ideal condition following constraints are imposed.

$\frac{1}{x} << 1$ or $\frac{Q\alpha\beta}{\gamma B_\gamma} << 1$ or $B_\gamma >> \frac{\alpha B}{\gamma} Q$

$\frac{2}{Qx} << 1$ or $\frac{2\alpha B}{\gamma B_\gamma} << 1$ or $B_\gamma >> \frac{\alpha B}{\gamma} 2$

$\frac{1}{Qx} << 1$ or $\frac{\alpha B}{\gamma B_\gamma} << 1$ or $B_\gamma >> \frac{\alpha B}{\gamma}$

So for the circuit to behave ideally the condition

$B_\gamma >> \frac{\alpha B}{\gamma} Q$ should be satisfied for $Q \geq 2$.

The exact error involved can be found out from the magnitude response of the expression (3.31). The program was run on the computer, for a given Q, ω_o and ~~x~~percentage of error; B_γ was found out. The precise results are given.

Q	Percentage error	x	$B_\gamma/2\pi$ f_o = 100 KHz	f_o=50 KHz
5	±5.136	16	16 MHz	8 MHz
10	±5	17	34 MHz	17 MHz
25	±5.165	19	25 MHz	47.5 MHz
50	±5	20	200 MHz	100 MHz

3.4 Two Op-amp Circuit

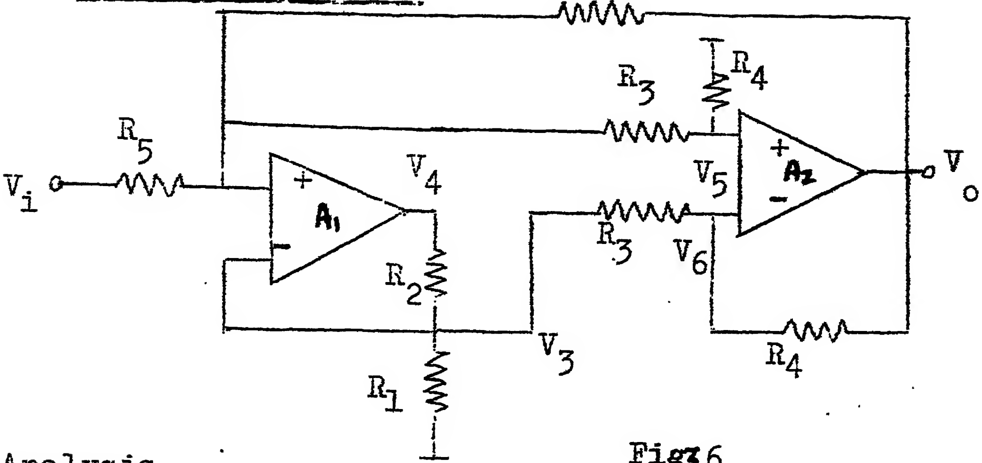


Fig3.6

Analysis

Let $\alpha = \dfrac{R_1}{R_1+R_2}$, $a = \dfrac{R_2}{R_3+R_4}$, $b = \dfrac{R_1}{R_3+R_4}$

$c = \dfrac{R_5}{R_3+R_4}$, $d = \dfrac{R_6}{R_3+R_4}$, $e = \dfrac{1}{(1/\alpha)+a}$

$f = \dfrac{1}{1+\frac{1}{a}+\frac{1}{b}}$, $\beta = \dfrac{R_3}{R_3+R_4}$, $\delta = \dfrac{R_5}{R_5+R_6}$

$m = \dfrac{1}{\frac{1}{1-\delta}+c}$, $n = \dfrac{1}{\frac{1}{\delta}+d}$

$$V_3 = \frac{V_4/R_2}{\frac{1}{R_1}+\frac{1}{R_2}+\frac{1}{R_3+R_4}} + \frac{V_o/(R_3+R_4)}{\frac{1}{R_1}+\frac{1}{R_2}+\frac{1}{R_3+R_4}}$$

$$= \frac{V_4}{\frac{\frac{R_1+R_2}{R_1}}{}+\frac{R_2}{R_3+R_4}} + \frac{V_o}{1+\frac{R_3R_4}{R_1}+\frac{R_3+R_4}{R_2}}$$

$$= \frac{V_4}{\frac{1}{\alpha}+1} + \frac{V_o}{1+\frac{1}{a}+\frac{1}{b}} \qquad (3.35)$$

$$= V_4e + V_of$$

$$V_4 = (V_2 - V_3) \frac{A_{o\alpha}}{1+s\tau_\alpha} \tag{1.36}$$

Substituting V_3 from (3.35) in (3.36), we have

$$V_4 = (V_2 - V_4\, e - V_o f) \frac{A_{o\alpha}}{1+s\tau_\alpha}$$

$$V_4\left(\frac{s\tau_\alpha}{A_{o\alpha}} + e\right) = V_2 - V_o f$$

$$V_4 = (V_2 - V_o f) \frac{B_\alpha}{s+eB_\alpha} \qquad (\text{if } A_{o\alpha} \gg 1) \tag{1.37}$$

Substituting (3.37) in (3.35),

$$V_3 = (V_2 - V_o f) \frac{eB_\alpha}{s+eB_\alpha} + V_o f$$

$$= V_2 \frac{eB_\alpha}{s+eB_\alpha} + V_o \frac{sf}{s+e\ B_\alpha}$$

$$V_5 = V_2 \frac{R_4}{R_3+R_4} = V_2(1-\beta)$$

$$V_6 = V_3 \frac{R_4}{R_3+R_4} + V_o \frac{R_3}{R_3+R_4} = V_3(1-\beta) + V_o\beta$$

$$V_o = (V_5 - V_6) \frac{A_{o\beta}}{1+s\tau_\beta}$$

$$= [V_2(1-\beta) - V_3(1-\beta) - V_o\beta] \frac{A_{o\beta}}{s\tau_\beta+1}$$

$$= [(V_2-V_3)(1-\beta) - V_o\beta] \frac{A_{o\beta}}{1+s\tau_\beta}$$

$$= (V_2-V_3)(1-\beta) \frac{B_\beta}{s+\beta B_\beta} \tag{3.39}$$

Substituting V_3 from (3.38) in (3.39)

$$V_o = \frac{\left[V_2 - V_3 \dfrac{eB_\alpha}{s+eB_\alpha} - V_o \dfrac{sf}{s+eB_\alpha}\right](1-\beta)B_\beta}{s+\beta B_\beta}$$

$$= \frac{\left[V_2 \dfrac{s}{s+eB_\alpha} - V_o \dfrac{sf}{s+eB_\alpha}\right](1-\beta)B_\beta}{s+\beta B_\beta}$$

$$= \frac{(V_2 - V_o f)s(1-\beta)B_\beta}{(s+eB_\alpha)(s+\beta B_\beta)} \qquad (3.40)$$

$$V_2 = \frac{V_i/R_5}{\dfrac{1}{R_5} + \dfrac{1}{R_6} + \dfrac{1}{R_3+R_4}} + \frac{V_o/R_6}{\dfrac{1}{R_5} + \dfrac{1}{R_6} + \dfrac{1}{R_3+R_4}}$$

$$= \frac{V_i}{\dfrac{R_5+R_6}{R_6} + \dfrac{R_5}{R_3+R_4}} \mp \frac{V_o}{\dfrac{R_5+R_6}{R_5} + \dfrac{R_6}{R_3+R_4}}$$

$$= \frac{V_i}{\dfrac{1}{1-\delta}+c} + \frac{V_o}{\dfrac{1}{\delta}+d}$$

$$= V_i m + V_o n \qquad (3.41)$$

Substituting V_2 from (3.41) in (3.40) we have

$$V_o = \frac{(V_i m + V_o n - V_o f)s(1-\beta)B_\beta}{(s+e\,B_\alpha)(s+\beta B_\beta)}$$

$$\frac{[V_i m + V_o(n-f)]s(1-\beta)B_\beta}{(s+eB_\alpha)(s+\beta B_\beta)}$$

$$V_o[(s+eB_\alpha)(s+\beta B_\beta) - (n-f)s(1-\beta)B_\beta] = V_i ms(1-\beta)B_\beta$$

$$\frac{V_o}{V_i} = \frac{s(1-\beta)mB_\beta}{s^2 s[eB_\alpha+\beta B_\beta - (n-f)(1-\beta)B_\beta]+e\beta B_\alpha B_\beta} \qquad (3.42)$$

$$\omega_o = \sqrt{(e\beta B_\alpha B_\beta)} \tag{3.43}$$

$$Q = \frac{\sqrt{(e\beta B_\alpha B_\beta)}}{eB_\alpha+\beta F_\beta - (x-f)(1-\beta)F_\beta} \tag{3.44}$$

$$G_o = \frac{(1-\beta)\ mB_\beta}{eB_\alpha+\beta F_\beta-(x-f)(1-\beta)F_\beta} \tag{3.45}$$

If $R_3+R_4 >> R_1 || R_2$, $a \simeq 0$, $b \simeq 0$, $e = \alpha$, $f = 0$

If $R_3+R_4 >> R_5 || R_6$, $c \simeq 0$, $d \simeq 0$, $m = 1-\delta$, $n = \delta$

Expression (3.42) reduces to

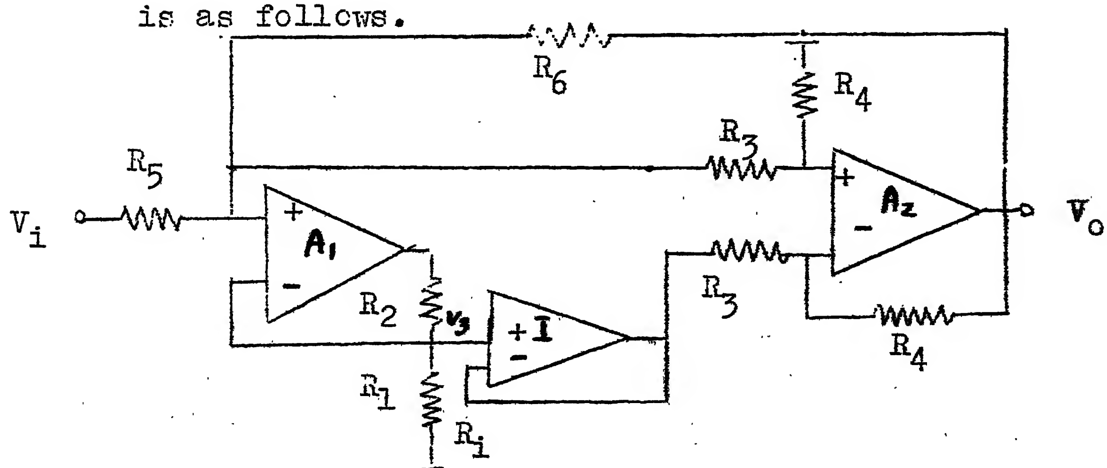
$$\frac{V_o}{V_i} = \frac{S(1-\beta)(1-\delta)B_\beta}{s^2+s[\alpha B_\alpha+\beta B_\beta - \delta(1-\beta)B_\beta]+\alpha\beta B_\alpha B_\beta} \tag{3.46}$$

$$\omega_o = \sqrt{(\alpha\beta B_\alpha B_\beta)} \tag{3.47}$$

$$Q = \frac{\omega_o}{\alpha B_\alpha+\beta B_\beta - \delta(1-\beta)B_\beta} \tag{3.48}$$

$$G_o = \frac{(1-\beta)(1-\delta)B_\beta}{\alpha B_\alpha+\beta B_\beta - \delta(1-\beta)B_\beta} \tag{3.49}$$

To avoid loading at V_3 a buffer may be put. The new circuit is as follows.



The buffer has to be flat over the range of frequency of operation. μA715 Op-amp which has bandwidth of 65 MHz is used as the buffer.

In this case,

$$a = \frac{R_2}{R_i}, \quad b = \frac{R_1}{R_i}$$

where R_i is the input resistance of the buffer which is very greater than R_2 and R_1. So $a \simeq 0$, $b \simeq 0$, $e = \alpha$, $f = 0$. The expression (3.42) reduces to

$$\frac{V_o}{V_i} = \frac{s(1-\beta)m\, B_\beta}{s^2+s[\alpha B_\alpha+\beta B_\beta - n(1-\beta)B_\beta] + \alpha B_\alpha\, \beta B_\beta} \tag{3.50}$$

$$\omega_o = \sqrt{(\alpha\beta B_\alpha B_\beta)} \tag{3.51}$$

$$Q = \frac{\omega_o}{\alpha B_\alpha + \beta B_\beta - n(1-\beta)B_\beta} \tag{3.52}$$

$$G_o = \frac{(1-\beta)mB_\beta}{\alpha B_\alpha + \beta B_\beta - n(1-\beta)B_\beta} \tag{3.53}$$

CHAPTER 4

DESIGN AND EXPERIMENTS

4.1.1 Three Op-amp circuit

To start with, the lowpass blocks were tuned to the desired centre frequency. Then the highpass block was realised by using one lowpass block and the differential summing blocks. The lowpass and the highpass blocks were cascaded and positive feedback was applied to obtain high Q. To realise a flat summer, the full bandwidth of the differential summing amplifier was used by keeping differential gain unity.

Tuning procedure

1. Adjust R_1/R_2 and R_3 by R_4 for tuning the centre frequency.
2. Adjust R_5/R_6 to obtain different Q.

4.1.2 Tuning of lowpass block

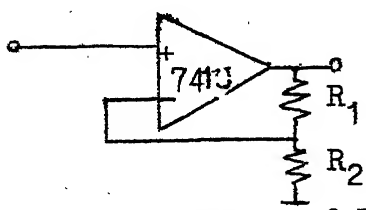


Fig.4.1: Lowpass Filter.

The electronic tuning of lowpass block was done by adjusting the resistor ratio R_1 and R_2. The low limit of R_1 is usually governed by the finite output resistance of the Opamps. The gain and phase response were measured and compared with the theoretical results as given in Table 4.1 and shown in graph 2.

4.1.3 Tuning of Highpass Block

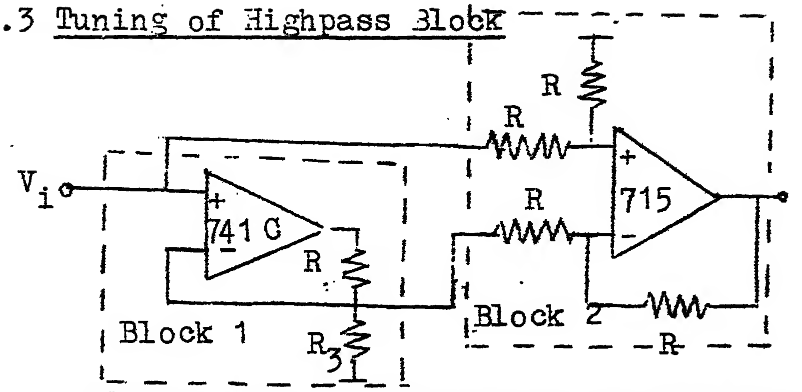


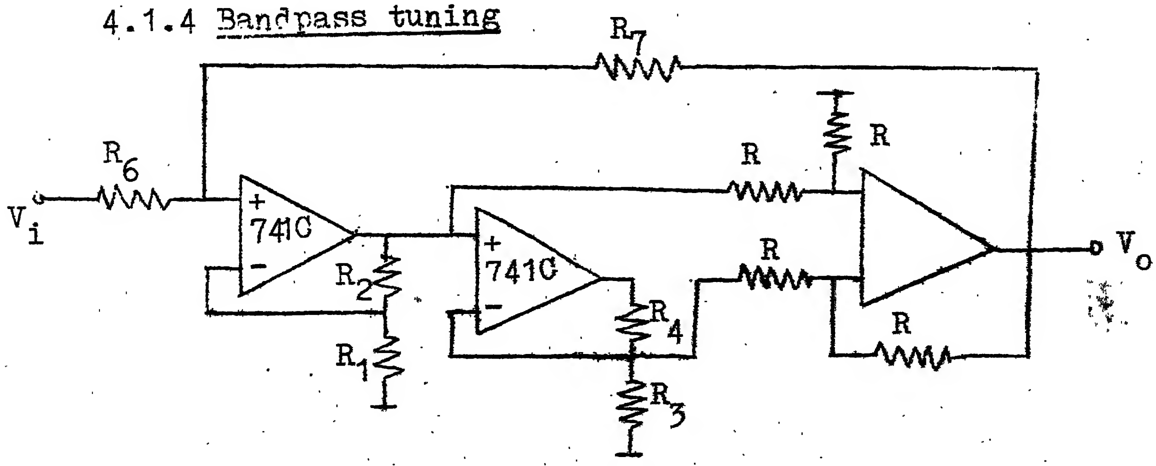
Fig.4.2: High-pass Filter

Block 1 is a lowpass filter and was tuned by the same method as before. Block 2 is a differential amplifier with unity gain. Its response both magnitude and phase with different values of R was measured and suitable R was selected so that it should not load the previous stage, i.e. block 1 as well as behave as a flat summer.

From the gain/phase reponse of the differential Op-amp in unity gain configuration as given in Tables 4.2 to 4.4 and shown in graph 3, we see that as the value of R increases, the peak frequency decreases (thereby bandwidth decreases) and the phase roll off also increases. We select R = 100K-ohm which satisfies the condition R R_3 R_4 to avoid loading.

Then the highpass response of the above circuit was measured and compared with theoretical results as given in Table 4.5 and shown in graph 3.

4.1.4 Bandpass tuning



The bandpass response with different feedback ratio, to obtain different Q, was measured experimentally and compared with the theoretical results as given in tables 4.6 to 4.8 and shown in graphs 4, 5 and 6.

4.1.5 Observations

$B_1 = 2\pi x833$ K rad/sec., $B_2 = 2\pi x833$ K rad./sec.

$R_1 = R_3 = 1$ K-ohm, $R_2 = R_4 = 7.3$ K-ohm, R = 100 K-ohm

$R_5 = 1$ K-ohm

R_6	Theoretical f_o(KHz)	Q	G_o	Experimental f_o(KHz)	Q	G_o
∞	100	0.5	4.2	100	0.5	4.36
5.1K-ohm	100	1.57	11.27	98	1.752	11.84
4.3K-ohm	100	4.0	26.6	96	4.44	27.27
3.5K-ohm	100	7.15	48.99	96	7.1	45.97

Comments:

1. From the bandpass response without the overall feedback, as shown in graph 3 we see that the phase is zero at 91 KHz and not at 100 KHz, where it should be. So when feedback was applied, the centre frequency shifted from 100 KHz to 96 KHz.

2. With higher Q, the signal at the non-inverting input of the first stage, increases, which increases the input signal level of the second stage and a point comes when Op-amps become saturated. This phenomenon is called the jump phenomenon. To avoid this, the test signal generator output has to be reduced considerably, with higher Q, to maintain linearity.

3. At low frequency of operation, the high gain of the Op-amp and at high frequency of operation, the slew rate of the Op-amp, limits the input signal level.

4.2.1 Two Op-amp circuit

Tuning of lowpass block: This tuning was done in the same manner as in the three Op-amp circuit.

4.2.2
Tuning of differential block: To start with, the inverting input was grounded, as shown in Fig. 4.4 and signal was applied at the non-inverting input. The resistor ratio R_3/R_4 was adjusted simultaneously for both the inputs and was tuned to the required frequency.

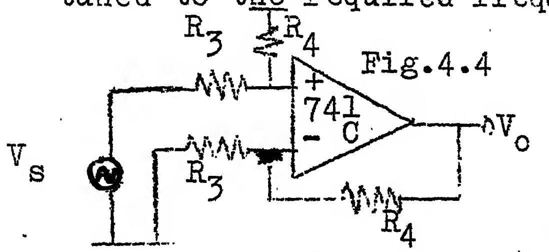


Fig.4.4

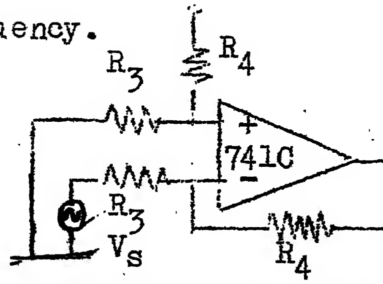


Fig.4.5

Similarly the process was repeated, with the signal at the inverting input and non-inverting input grounded as shown in Fig.4.5.

4.2.3
Tuning of bandpass block: The circuit was made as shown in Fig. 4.6. The bandpass response at different feedback ratio to obtain different Q was measured experimentally and compared with the throretical resusts as given in Tables 4.9 to 4.13 and shown in graphs 7 and 8.

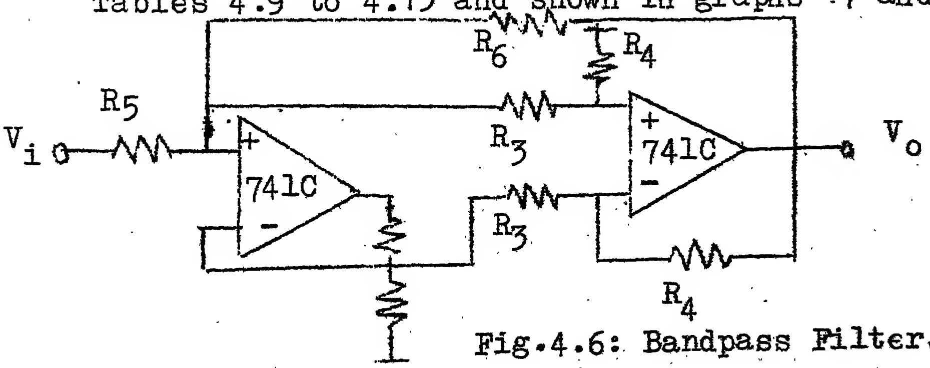


Fig.4.6: Bandpass Filter.

Observations (refer Tables 4.9 to 4.11)

$B_1 = 2\pi \times 833$ K rad./sec., $B_2 = 2\pi x833$ K rad./sec.

$R_1 = 200$ ohm, $R_2 = 3.15$ K-ohm $R_3 = 6.33$ K-ohm

$R_4 = 100$ K-ohm, $R_5 = 1$ K-ohm.

	Theoretical			Experimental		
R_6	f_o(KHz)	Q	G_o	f_o(KHz)	Q	G_o
8 K-ohm	50.0	3.12	52.96	47.5	4.15	55.55
7.5K-ohm	50.0	9.5	98.73	47.5	7.15	97.08
7.4K-ohm	50.0	8.45	116.768	47.2	11.85	111.111

Comments

From the experimental results, it was seen that the frequency shifts from theoretical value of 50 KHz to 47.5 KHz and accordingly the Q factor changes. This was due to loading of the first stage by the second stage. So the value of R_1 was reduced. Moreover the value of R_4 and R_3 was reduced which was causing excess phase shift in the differential stage.

Observations (refer Tables 4.12 to 4.15)

$B_1 = 2\pi x833$ K rad./sec., $B_3 = 2\pi x900$ K rad./sec.

$R_1 = 35$ ohm, $R_2 = 450$ ohm, $R_3 = 1.94$ K-ohm,

$R_4 = 33$ K-ohm, $R_5 = 1$ K-ohm.

In tuning the first stage lowpass filter the output resistance of the Op-amp has been taken into consideration.

R_6	Theoretical f_o(KHz)	Q	G_o	Experimental f_o(KHz)	Q	G_o
10 K-ohm	50	1.95	29.75	50	1.953	28.74
8.6 K-ohm	50	3.59	51.14	49.2	3.42	47.62
8.1 K-ohm	50	4.8	75.5	49.2	5.23	62.5
7.6 K-ohm	50	11.36	166.24	49.2	8.33	100.0

Comments

From the experimental result, it was seen that as the feedback was increased, the gain was reduced. This was due to offset at the output of Op-amps. So the offset was made null. To avoid the effect of output resistance of the first stage Op-amp, the value of R_1 was increased. To avoid the loading of first stage, a buffer was put as shown in Fig. 4.7. μA715 which has a bandwidth of 65 MHz, was used as a buffer.

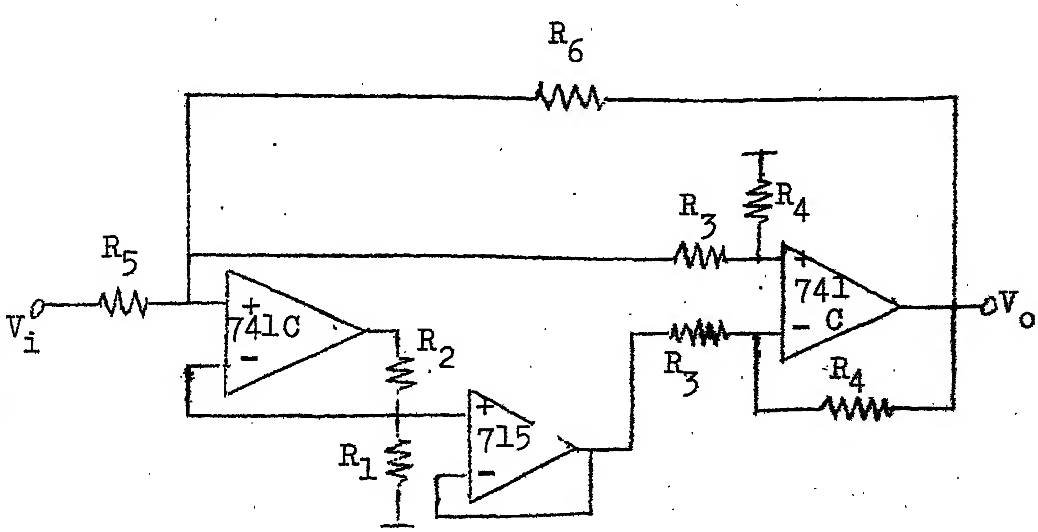


Fig.4.7: Bandpass filter with buffer 715.

Observations (50 KHz Tuning): (refer Tables 4.16 and 4.17 and graph 7).

$B_1 = 2\pi x833$ K rad./sec., $B_3 = 2\pi x900$ K rad./sec.

$R_1 = 2$ K-ohm, $R_2 = 31.5$ K-ohm, $R_3 = 1.94$ K-ohm

$R_4 = 33$ K-ohm, $R_5 = 2$ K-ohm.

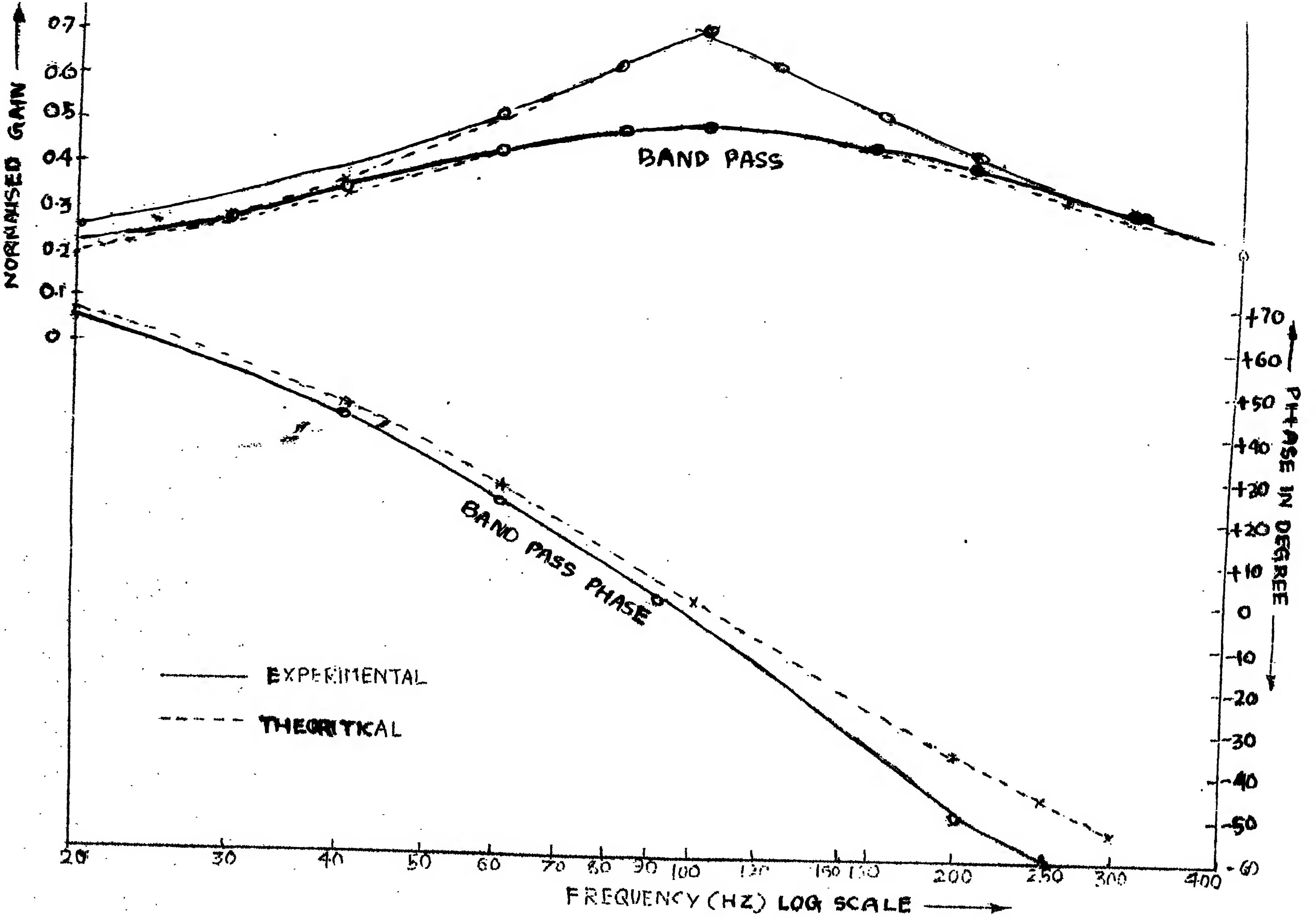
	Theoretical			Experimental		
R_6	f_o(KHz)	Q	G_o	f_o(KHz)	Q	G_o
15 K-ohm	50	10	150.37	49.2	10	149.1
16.2K-ohm	50	4.35	65.12	49.2	4.4	65.00

100 KHz Tuning: (refer Table 4.18 to 4.20 and graph 8).

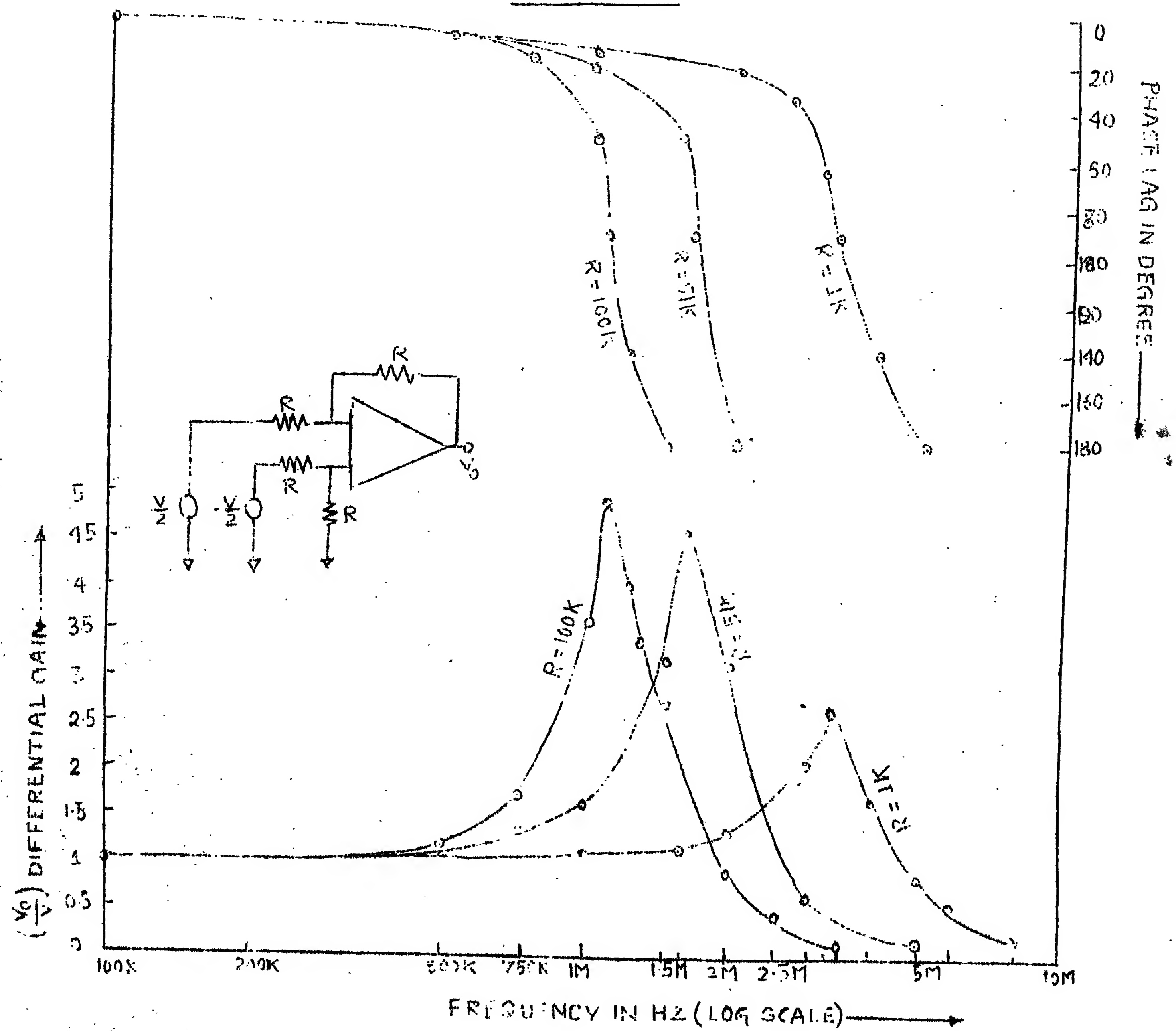
$R_1 = 2$ K-ohm, $R_2 = 14.66$ K-ohm, $R_3 = 4.125$ K-ohm

$R_4 = 33$ K-ohm, $R_5 = 2$ K-ohm

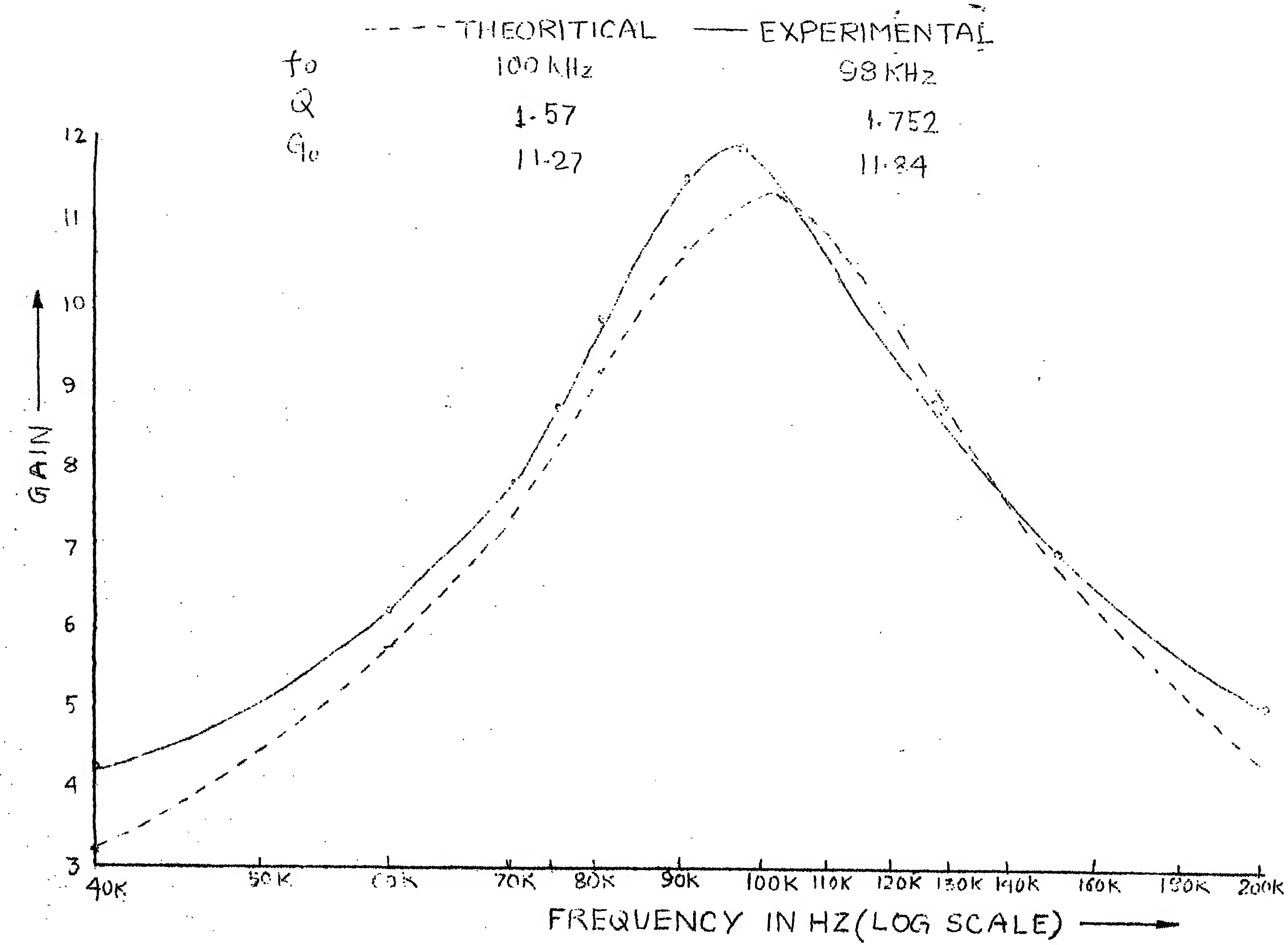
	Theoretical			Experimental		
R_6	f_o(KHz)	Q	G_o	f_o(KHz)	Q	G_o
6.3 K-ohm	100	6.756	39.2	99.2	6.74	38.46
6.1 K-ohm	100	9.89	55.7	99.2	9.78	54.82
6 K-ohm	100	13.0	74.24	99.2	13.0	72.78


NORMALISED GAIN
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0
BAND PASS
BAND PASS PHASE
EXPERIMENTAL
THEORITICAL
PHASE IN DEGREE
+70
+60
+50
+40
0
+10
-10
-20
-30
-40
-50
-60
20
30
40
50
60
70
80
90
100
200
250
300
400
FREQUENCY (HZ) LOG SCALE

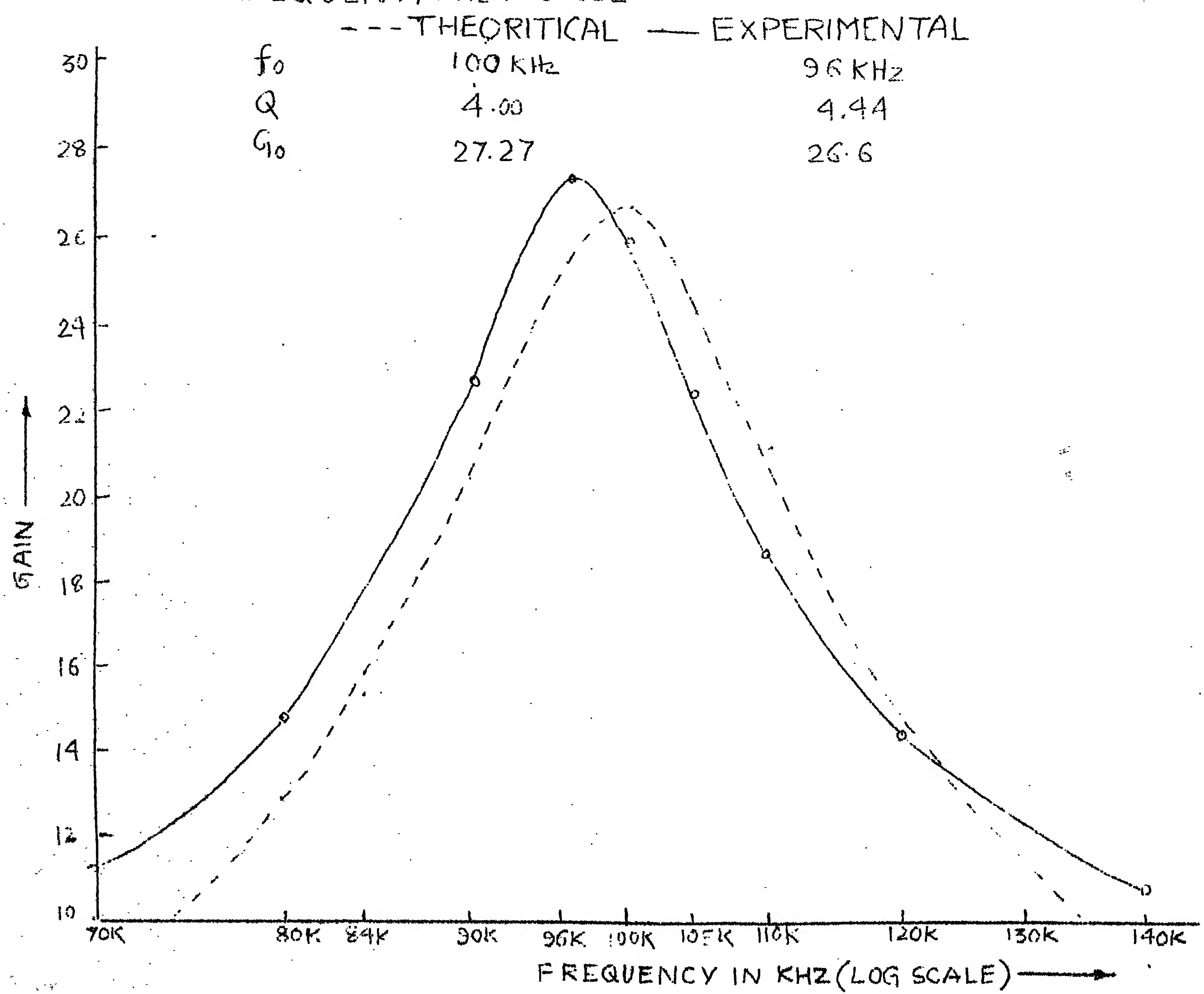
GRAPH - III



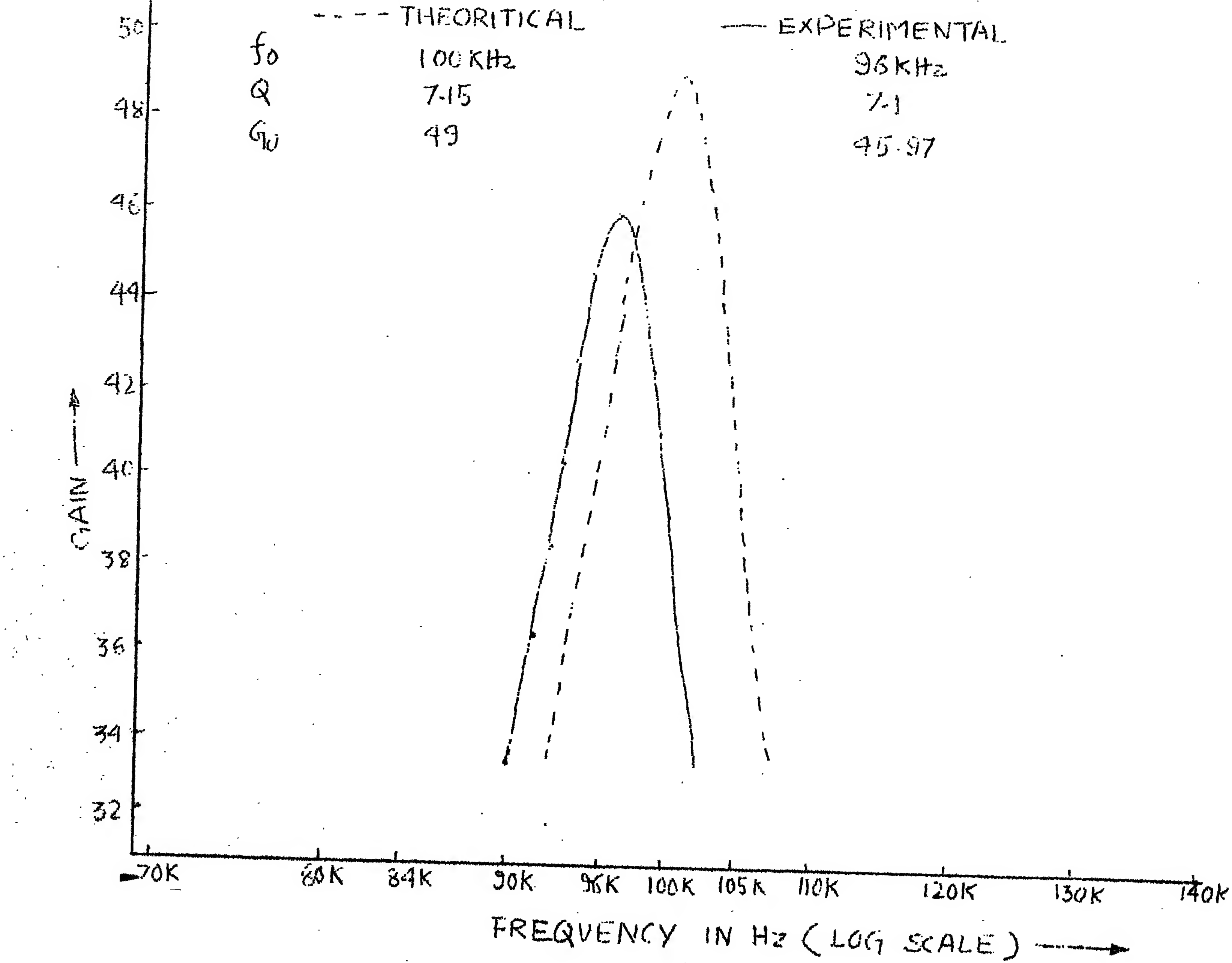
PHASE LAG IN DEGREE
0
20
40
60
80
100
120
140
160
180
R = 100K
R = 51K
R = 1K
$\left(\frac{V_o}{V}\right)$ DIFFERENTIAL GAIN
5
4.5
4
3.5
3
2.5
2
1.5
1
0.5
0
FREQUENCY IN Hz (LOG SCALE)
100K
200K
500K
750K
1M
1.5M
2M
2.5M
5M
10M
R
R
R
R
$\frac{V}{2}$
$-\frac{V}{2}$
V_o



THEORITICAL
EXPERIMENTAL
fo
Q
Go
100 kHz
1.57
11.27
98 KHz
1.752
11.24
GAIN
12
11
10
9
8
7
6
5
4
3
40K
50K
60K
70K
80K
90K
100K
110K
120K
130K
140K
160K
180K
200K
FREQUENCY IN HZ(LOG SCALE)



--- THEORITICAL — EXPERIMENTAL
f0 100 KHz 96 KHz
Q 4.00 4.44
G10 27.27 26.6
GAIN
FREQUENCY IN KHZ (LOG SCALE)
70K 80K 84K 90K 96K 100K 110K 120K 130K 140K
10 12 14 16 18 20 22 24 26 28 30



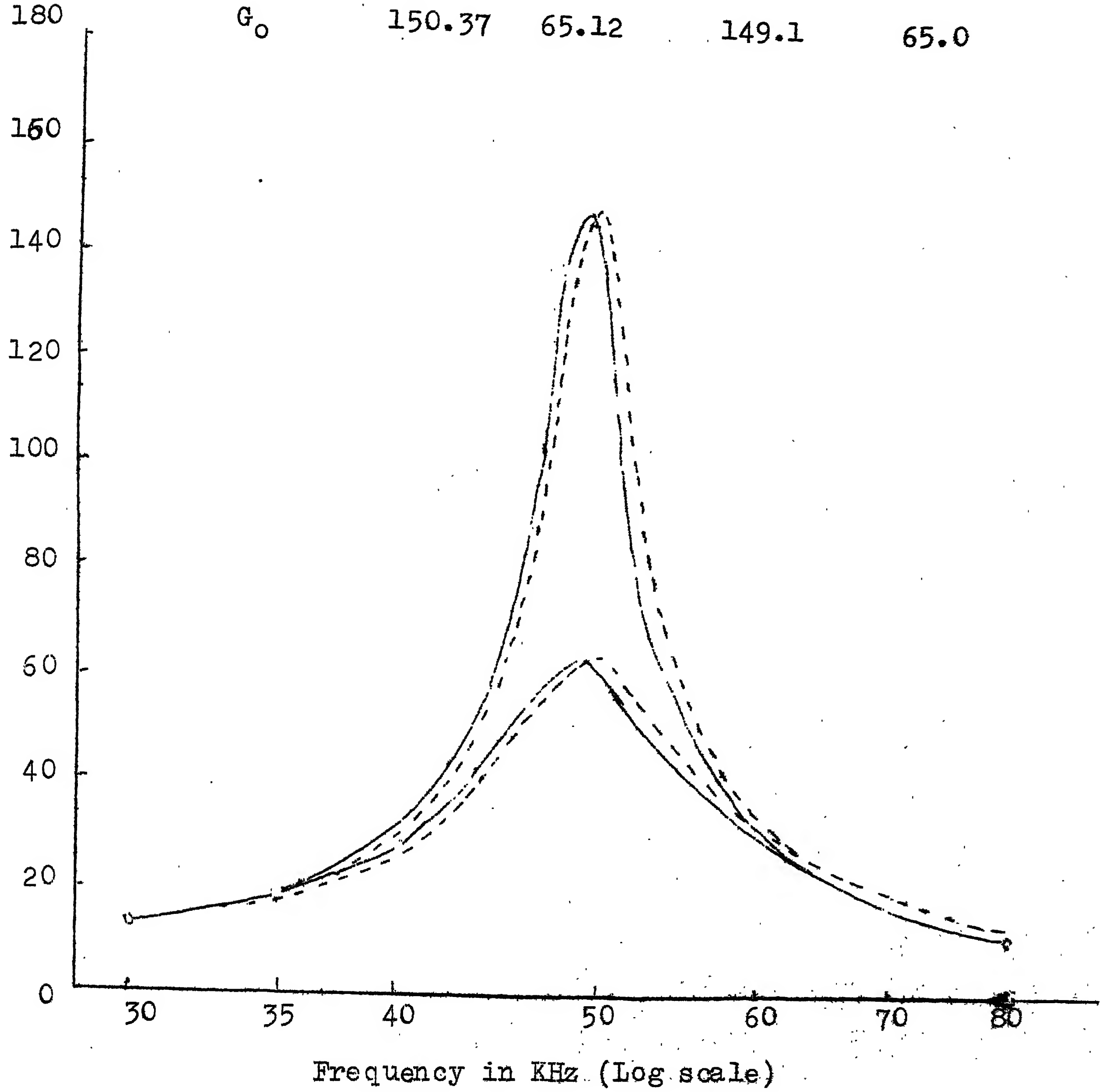
- - - - THEORITICAL
—— EXPERIMENTAL
fo
Q
Gu
100 KHz
7.15
49
96 KHz
7.1
45.97
GAIN
50
48
46
44
42
40
38
36
34
32
70K
80K
84K
90K
95K
100K
105K
110K
120K
130K
140K
FREQUENCY IN Hz (LOG SCALE)

GRAPH VII

FREQUENCY RESPONSE OF BANDPASS FILTER

----THEORETICAL ———EXPERIMENTAL

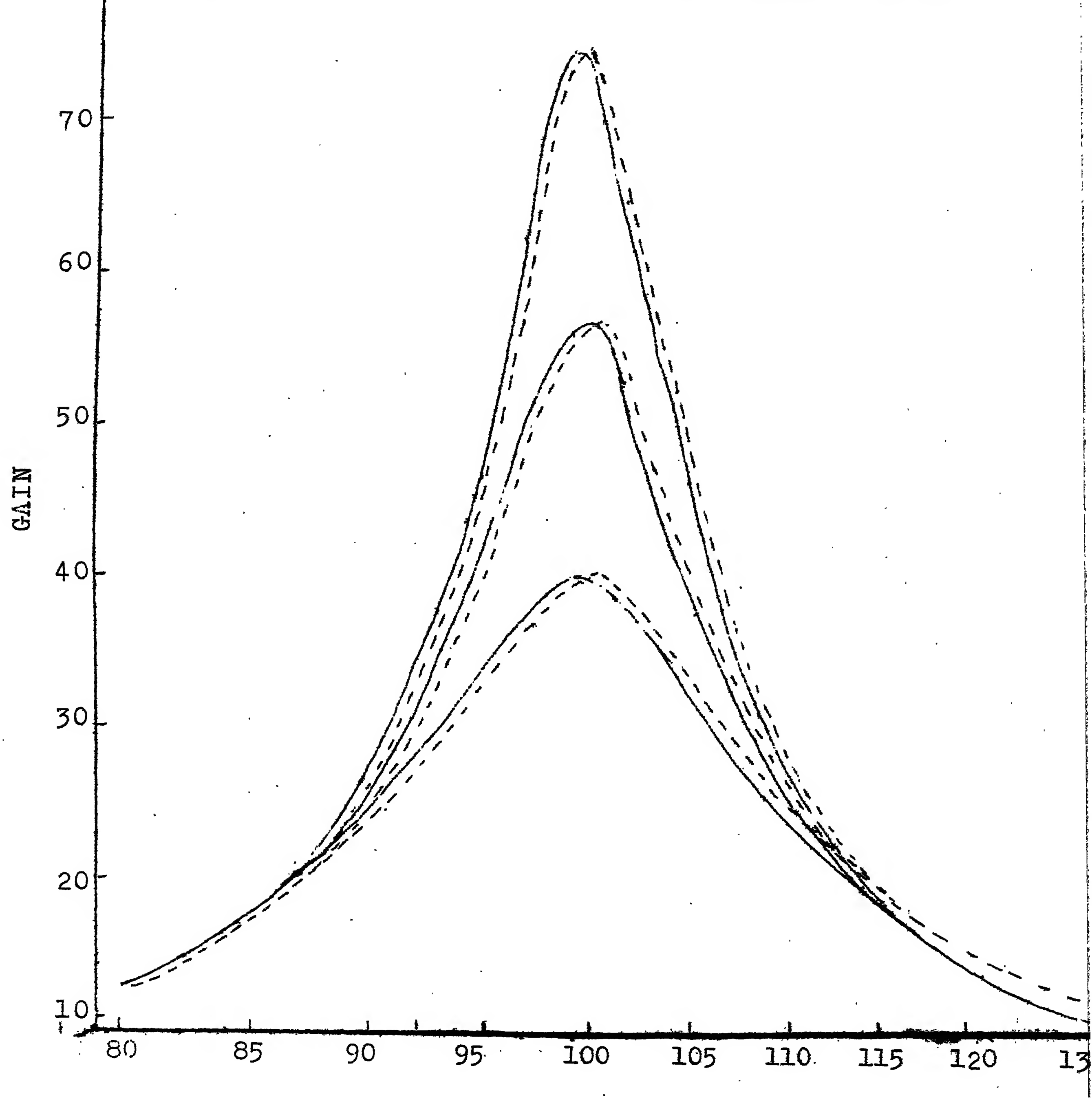
	THEORETICAL I	THEORETICAL II	EXPERIMENTAL I	EXPERIMENTAL II
f_o(KHz)	50	50	49.2	49.2
Q	10	4.35	10.0	4.4
G_o	150.37	65.12	149.1	65.0



FREQUENCY RESPONSE OF BANDPASS FILTER

----- THEORETICAL ——— EXPERIMENTAL

	Theoretical I	Theoretical II	Theoretical III	Experimental I	Experimental II	Experimental III
f_o (KHz)	100	100	100	99.2	99.2	99.2
Q	13	9.89	6.756	13.0	9.78	6.74
G_o	74.24	55.70	39.20	72.78	54.82	38.46



CHAPTER 5

CONCLUSION

5.1 Circuit Evaluation

In this thesis, active R bandpass filter, using three Op-amps and two Op-amps separately, has been proposed. Since these circuits use the single pole Op-amp model, these circuits can be used at high frequencies without the drawback encountered in the conventional circuit design. Moreover, since no external capacitors are required, these circuits are suitable for integration.

Op-amp 741C has been experimentally modelled and it is seen from Chapter 2 that the experimental model matches closely with the throretical model, till the 0 dB frequency. But the phase response deviates from the theoretical one pole model from 100 KHz onwards. This is due to the effect of 2nd pole of the Op-amp. This excess phase shifts limits the performances of 741C to 100 KHz in filter circuits.

So the single pole model namely $A(s) = A_0/(s+1)$ has to be taken into design consideration when operating at and below 100 KHz. The second pole model namely $A(s) = A_0/((1+s)(\frac{1-S/2\omega_2}{1+S/2\omega_2}))$ has to be considered when operating in the range between 100 KHz and 1 MHz.

The proposed three Op-amps bandpass filter was developed and the design equations were derived in Chapter 3. The involved

error calculation due to the non-ideality of the summing amplifier was carried out. A program was run on the amplifier for a given Q, ω_o and percentage of error, the bandwidth of the summing amplifier was obtained. The circuit was tested at f_o = 100 KHz at different Q. The experimental and theoretical results were compared as shown in the graphs 4,5 and 6.

Then the two Op-amps circuit was developed and design equations were derived. The circuit was tested at f_o=50 KHz and f_o = 100 KHz at different Q. The experimental and the theoretical results were compared as shown in the graphs 7 & 8.

The proposed bandpass filters can be made oscillatory since a negative term appears in the S term of the characteristic equations. In the three Op-amp circuit at V_3 we can get lowpass response whose tranfer function is given by

$$T_{LP} = \frac{V_3}{V_i} = \frac{(1-\delta)\ B^2}{s^2+s\ \ B(2-\frac{\delta}{\alpha})+(\alpha B)^2}$$

5.2 <u>Problems and Applications of Active R Filter:</u>

In practice the characteristics of inexpensive hybrid or monolithic amplifier are really not too good. The open loop frequency response of 741C matches closely both in magnitude and phase to the assumed integrator model for the operational amplifier. The input and output impedance versus frequency characteristics of these Op-amps show that their effect on filter performances is a second order offset when compared to the open loop frequency response. Deterioration of common mode rejection ratio when both the inputs of the Op-amps are used effects the filter performances.

Slewrate limitations, on the other hand are as severe a limitation as the open loop frequency response. The best solution to this problem is to select an Op-amp with better slewrate.

The jump phenomenon in the filters due to the slewing-rate of the amplifiers is well known. The maximum output voltage swing for different frequencies is to be kept lower than the values shown below.

f_o (KHz)	1	5	10	50	100	175
V_{max}(volt)	15	15	5	1	0.5	0.275

Since d.c. coupling is involved, the output offset of the Op-amps drastically effects the pole location and gain of the filter. So the Op-amps are to be modelled and operated at zero offset conditions.

The limiting factor for the tuning range both of f_o and Q depends on the validity of the single pole model of the amplifiers used. For 741C this point is about 100 KHz, above this frequency the phase shift of the amplifier becomes more than 90° due to the phase contribution by second pole.

Stabilising the circuit performances under varying temperature and bias voltage can be achieved by employing suitable compensation methods using JFET and thermistors.

Whereas most of the early active RC filters are designed to be essentially independent of the amplifier parameters, assuming only that the gain is large and constant, in active R

filter, the frequency response of the circuit is derived from the amplifier roll off characteristics and so the filter performances depend therefore on the amplifier parameters and in particular on the gain bandwidth product B. This necessi the knowledge of not only the values of B, but also of the variation of B, with temperature and supply voltage etc. In the proposed filter circuit, ω_0 is proportional to $\sqrt{\alpha\beta B_\alpha B_\beta}$. This suggests a method of controlling ω_0 by varying the bias current of the Op-amps.

The author is of the opinion that Active R filters are practical using current Op-amp technology. However, in the future, it would seem reasonable to assume that Op-amps specifically designed for active R filters could be constructed with optimum slew rate, bandwidth and frequency roll off properties of active R filter design. Such design would anticipate the use of all available bandwidth at all frequency upto the unity gain frequency and thus would balance the parameters of the Op-amps in such a way to achieve a good compromis between slewrate, bandwidth and roll off. With such an advance technology, it might be feasible in the future to design active R filters at microwave frequencies.

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